A Low-Cost Near-Constant Impedance Signal Mixer for Analogue Synthesizers

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Abstract

Almost every analogue synthesizer contains several signal mixers, both for audio signals and for control voltages. The traditional inverting summing mixer has several drawbacks relating to circuit noise, input impedance and DC offsets. This paper presents a new view of an old technique that solves several of these issues. It analyses the theoretical performance of the old and new circuits, and presents both a theoretical example, and experimental results that confirm the theory.

1 Introduction

Almost every analogue synthesizer contains several signal mixers, both for audio signals and for control voltages (CV). The most common structure for these mixers is the well-known inverting summing op-amp circuit. It has proven popular over the years, due in no small part to its simplicity, ease of understanding and excellent channel isolation. Nonetheless it does suffer from circuit characteristics which require the designer to make compromises in circuit noise, input impedance, and DC offsets.

This paper presents a new approach to this classic circuit, which addresses some of the shortcomings of the original, yet using the same set of components. Experimental results show that the new arrangement is at least comparable and mostly better than the existing circuit.

This paper is structured as follows. First the existing circuit is presented and analysed to show its major shortcomings. Then the the new circuit is presented, with the same circuit characteristics examined as for the classic circuit, comparing the two. Experimental results are shown to confirm the new circuit's behaviour.

2 What's Wrong With The Old Circuit?

The most common input mixer circuit comprises a potentiometer (pot), either linear or logarithmic law, with the wiper driving an inverting op-amp, usually with unity gain. A two-channel mixer is shown in Figure 1 below.

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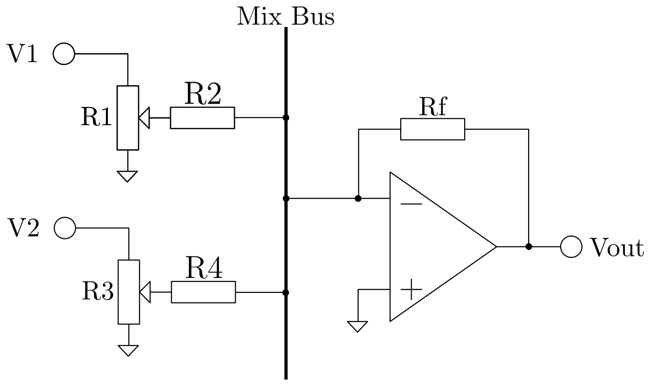

Figure 1: *Typical 2-input mixer. This can be extended to three or more inputs by adding a pot and summing resistor per input to the mix bus.*

A proportion (k, where k lies between 0...1) of the input signal V_1 is tapped from the pot R_1 by the wiper, and fed into the input resistor, R_2, of an inverting amplifier. The feedback resistor, R_f, provides a feedback path from the op-amp's output back to the input. The gain applied to V_1 is simply $-\frac{R_f}{R_2}$. The same behaviour applies to V_2, R_3 and R_4, where the gain applied to V_2 is $-\frac{R_f}{R_4}$.

The action of the op-amp is to drive the output to keep the *-ve* input the same as its *+ve* input. This forces the common *mix bus* to what is called a *virtual earth.* This gives excellent channel separation[1].

2.1 Input Resistance

To avoid undue loading of signal sources the input resistance seen at the mixer inputs should be much larger, say 10×, than that of the source[2]. Also, where a single signal source drives multiple inputs then the input resistances must be constant, so that where there are loading effects any error so introduced is independent of the setting of the input attenuators (R_1 and R_3 in Figure 1).

In the classic circuit the input resistance can vary quite substantially. It depends on the values of R_1 and R_2, and of the setting, k, of the pot:

$$R_{in} = (1 - k)R_1 + (R_2 \parallel kR_1)$$

When the wiper is at the grounded end of the track the input sees only the resistance of R_1. However, as the wiper moves towards the other end of the track the effect of the loading of the wiper by R_2 reduces the input resistance. In a typical case where both

[1] More advanced techniques such as balanced mix busses or grounded busses is beyond the scope of this paper.

[2] Early audio systems had 600Ω outputs, requiring a minimum of 6k input resistance. This in turn lead to the common 10k input resistance (nearest preferred value). In later years input resistances have crept up to around 47k or more.

resistors have the same value the input resistance will vary over a 2:1 range. Figure 2 shows this for 100k resistors.

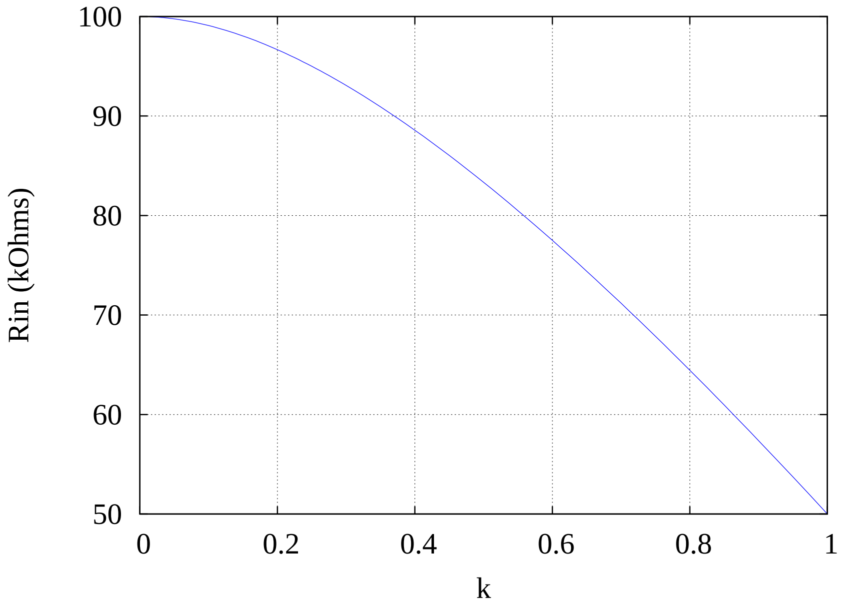

Figure 2: *The input resistance of the classic op-amp mixer input for a 100k pot and 100k summing resistor. The pot position is k, 0 being fully-CCW, 1 being fully CW. Note that the input resistance varies over a 2:1 range.*

Where the signal source has non-zero impedance, typically 300-1k, and where the source is driving multiple inputs in parallel, any changes in the input resistance will lead to changes in the voltage seen by all the mixer inputs. For control voltages controlling pitch this would be audibly noticable.

2.2 Resistance Seen By Op-Amp

The *-ve* input pin of the op-amp would ideally "see" a zero source resistance. Then the input bias current, I_{IB}, would introduce no offset. If this is not possible, then the next best solution, especially for CV mixing where DC accuracy is important, is a constant resistance so that any offset, once trimmed, remains trimmed.

Consider the upper channel in Figure 1. The *-ve* pin sees R_f in parallel with R_2 in series with the upper and lower sections of R_1 in parallel (assuming the source has negligible resistance). The value of R_{-ve} is given by

$$R_{-ve} = R_f \parallel \Big(R_2 + (kR_1 \parallel (1-k)R_1)\Big)$$

As shown in Figure 3 the lowest value of R_{-ve} is when the wiper is at either end of R_1 (assuming V_{in} is ideal), the highest when the wiper is exactly half-way.

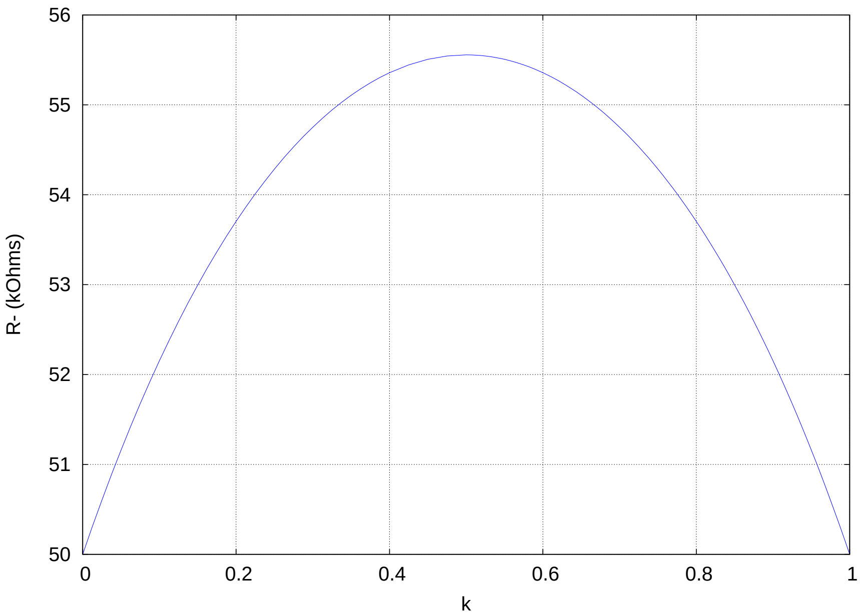

Figure 3: *The resistance seen by the -ve input of the op-amp for a 100k pot and 100k summing resistor. It varies by* 11%*. Any offset due to bias currents would vary to the same extent due to Ohm's Law.*

2.3 Conformance to Ideal Behaviour

The loading effect that R_2 has on the wiper of R_1 introduces an error[3]. For example, when the wiper is half-way ($k = 0.5$), then R_2 is in parallel with the bottom half of R_1. Thus the voltage at the wiper is not $0.5V_{in}$ but somewhat less – exactly how much less depends on the value of R_2.

This affects both linear and logarithmic law potentiometers. For a linear pot in a CV mixer the user will find that the control turned halfway does not equate to a 50% input level, but somewhat less. For a logarithmic pot in an audio mixer the user may not notice a change in the control law, unless the value of R_2 is significantly less than R_1.

Analysis of the classic circuit shows that the transfer function is

$$\frac{V_{out}}{V_{in}} = \frac{-kR_f}{R_2 + (k - k^2)R_1}$$

An example of the deviation from the ideal is shown in Figure 4. This is for a typical situation where all three resistances are 100k. To improve this the ratio between R_1 and

[3]Sometimes this "bending" of the pot's control law is desirable. For example some audio designers prefer to use linear pots and then bend the law into something approximating a log-law pot with a wiper resistor much smaller (say, one-sixth) than the track resistance. The merits or otherwise of this technique are not discussed in this paper.

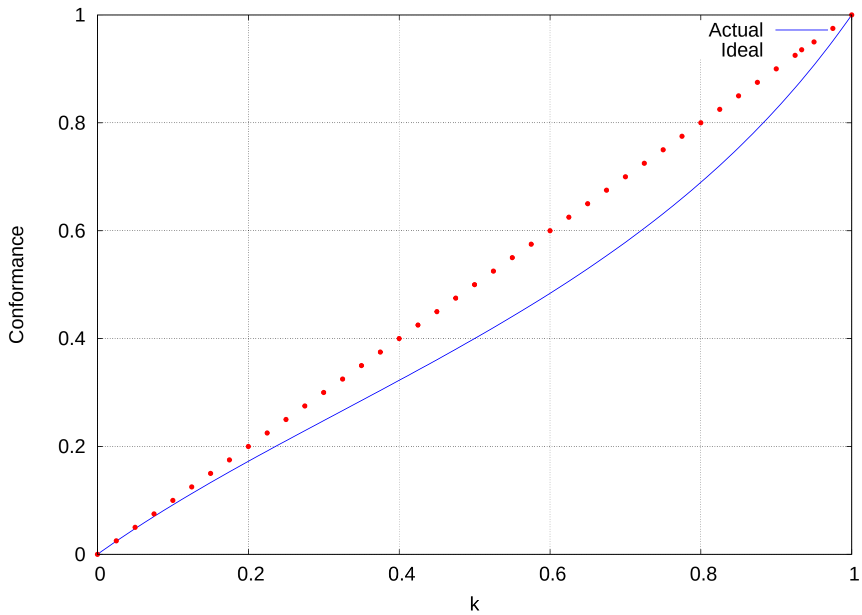

Figure 4: *The conformance to the pot law is clearly visible as the transfer function deviates from the ideal straight-line. All resistors and the pot are 100k.*

R_2 must be improved; *e.g.*, reducing the pot's resistance to, say, 47k which will reduce the loading effect that R_2 has.

2.4 Noise Annoys

The generally-applied rule for pot wiper loading is to make the wiper load resistance at least 10× that of the track resistance. So, for example, for R_1 of 100k we would need to set R_2 to a minimum of 1M0, and also for R_f. But this would introduce large DC offsets due to I_{IB}, and undesirable levels of thermal noise[4] for audio mixers.

3 An Improved Signal Mixer

Can anything be done to improve the behaviour of the classic signal mixer, as described above, without incurring higher component count and cost? For example, we could insert a unity-gain buffer on each input prior to the attenuator. But this adds considerable cost, especially if there are many inputs.

The answer is a clear *yes*. This section describes a solution, shown in Figure 5, which addresses several of the major shortcomings of the classic signal mixer, yet without any

[4]Thermal, or Johnson-Nyquist, noise voltage of a resistor is given by $v_n = \sqrt{4kTBR}$, where k is Boltzmann's constant, T is the device temperature, B is the bandwidth over which the voltage is measured, and R is the resistance. For example, a 100k resistor measured over a 22kHz bandwidth at room temperature produces a thermal noise voltage of $6\mu V$.

additional cost or complexity.

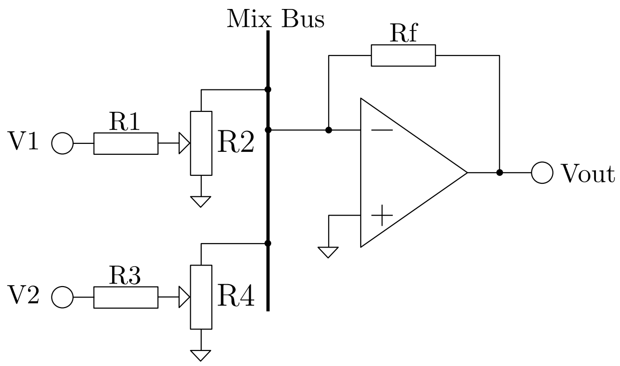

Figure 5: *The proposed new signal mixer design. Note especially the connection to the wiper. Additional inputs connect to the mix bus in the same way as R_1,R_2.*

At first glance this appears very *wrong*. The signal is going *into* the pot's wiper, whereas the majority of uses of pots see the signal coming from the wiper. When the wiper is at the grounded end, the signal goes to ground. Well, yes, through R_1. And when the wiper is at the top it starts to look like an ordinary inverting amplifier, with R_1 as the bus summing resistor, but then R_2 is sitting between the mix bus and ground. But since the mix bus is already a virtual earth no current flows through it.

But the change in the order of R_1 and R_2 has an important effect: the input resistance seen by the source is now dominated by R_1. The role of R_2 is now to share current between ground and the mix bus. One way to think of it is this:

- In the classic circuit, the pot's resistance sets the input resistance and the wiper resistor should be many times larger (ideally infinite);
- In the new circuit, the wiper resistor sets the input resistance, and the pot resistance would ideally be zero.

In practice, we can more easily (and preferably, due to noise and DC offsets) approximate a low resistance than we can a high resistance.

3.1 Input Resistance

The op-amp ensures the right-hand side of R_2 is at a virtual earth, which helps simplify the input resistance calculation.

$$R_{in} = R1 + (kR_2 \parallel (1-k)R_2)$$

Rearranging,

$$R_{in} = R1 + (k - k^2)R_2$$

While this may look no better than the classic case, note that the input resistance is dominated by R_1 (the maximum value of $k - k^2$ is 0.25). We can therefore set R_1 to the desired high input resistance, and set R_2 to a lower value thus minimising changes to the input resistance (Figure 6).

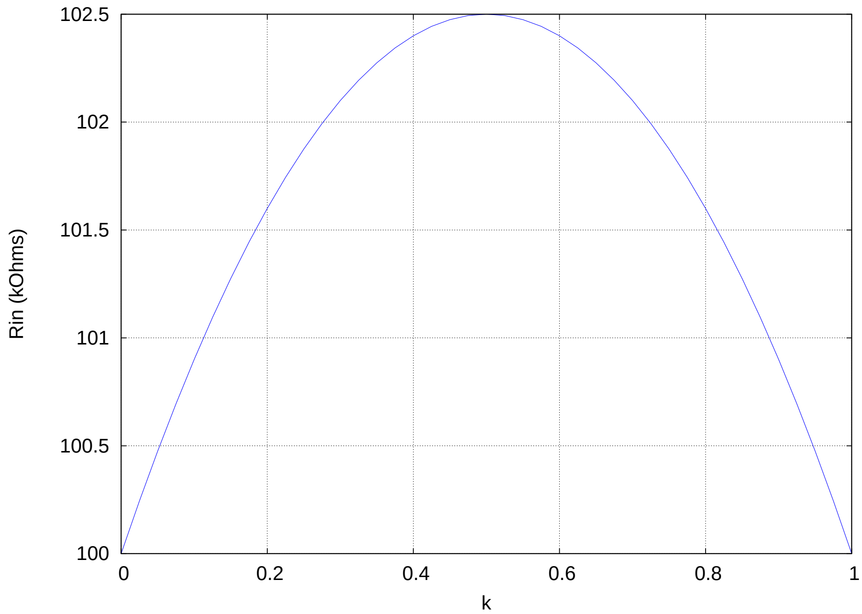

Figure 6: *The input resistance of the new op-amp mixer input for a 100k input resistor and 10k pot. The deviation is only* 2.5%*. To achieve a similar result from the classic circuit the summing resistor would need to be about 4M0.*

3.2 Resistance Seen By Op-Amp

Given that we can now use a small value for R_2 this tends to dominate the parallel resistance network. For unity gain we make R_f the same as R_1, which would typically be of the order of 50k to 100k. With R_2 set to, say, 10k the resistance seen by the *-ve* pin is going to be less than 10k.

This benefits the DC operation as the effects of any input bias current are reduced. And while it may vary, it is going to be an order of magnitude or so smaller than for the classic case.

3.3 Conformance to Ideal Behaviour

Surprisingly this new circuit exhibits the same theoretical conformance to ideal behaviour as the classic circuit. However the fact that we can set R_2 to a much lower value can be used to minimise the deviation from ideal (shown in Figure 7), which is not feasible with the classic circuit.

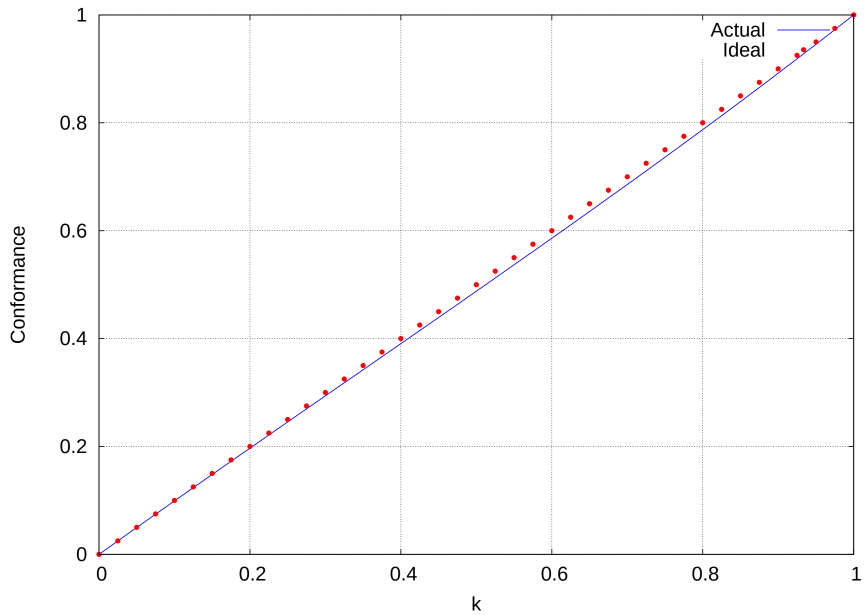

Figure 7: *The new circuit is much closer to the ideal pot law.*

Plotting both old and new conformance curves shows just how markedly different and improved the new circuit is over the old circuit.

3.4 Noise Analysis

The new circuit proposed in this paper solves many problems present in the classic circuit. However it is by no means perfect, and there are two important issues which need further exploration.

3.4.1 Noise Gain is Higher

Noise Gain is the gain applied to the inherent noise generated by the op-amp itself. It is defined as

$$\text{Noise Gain} = 1 + \frac{R_f}{R_{in}}$$

To consider the worst-case Noise Gain we calculate the lowest value of R_{in} for N inputs (assuming they are all the same). For the classic circuit the Noise Gain is

$$\text{Noise Gain} = 1 + \frac{NR_f}{R_{in}}$$

For example, with R_f and R_{in} both 100k and for $N = 3$ inputs the Noise Gain is 12dB.

The new circuit exhibits a much lower value of R_{in} due to the placement of the low-valued pots.

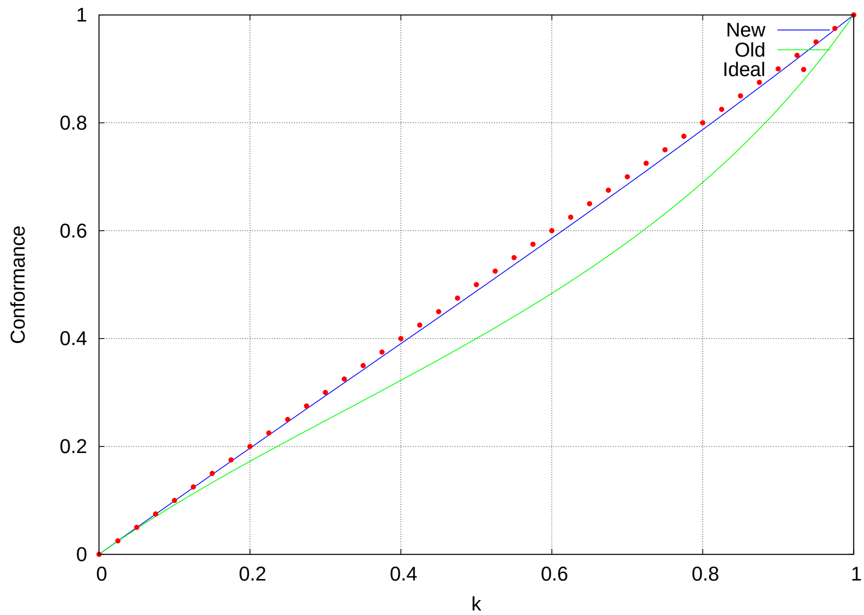

Figure 8: *The new circuit conforms much closer to the ideal than the old, primarily due to the smaller value of R_2.*

$$\text{Noise Gain} = 1 + \frac{N R_f}{R_{pot} \parallel R_{in}}$$

Continuing the earlier example (R_f and R_{in} are 100k, R_{pot} is 10k, and there are three inputs) the Noise Gain is 30.36dB. This is 18dB (8×) greater than for the classic circuit. This will result in larger DC offsets (due to V_{OS}) and general noise (due to e_n and/or i_n). While at first this may seem problematic, it must be weighed against the benefits that the new circuit arrangement brings (better law conformance, improved input resistance behaviour, *etc*).

3.4.2 Summing Bus is Large

The original circuit separates the pot from the summing bus through the wiper resistors. This arrangement allows for the summing bus to kept away from the front panel.

The new circuit, however, brings the summing bus to one end of each pot. This can lead to greater capacitive coupling of noise into the summing bus. However, in practice such a summing bus would be limited to a small area, perhaps a few centimetres in length. Contrast that against a mixing desk with multiple summing busses spanning a metre or more. Given that body of examples it is hard to argue against the new circuit on that basis alone.

4 A Paper Example

To show the benefits of this new mixer circuit we now present an example application, using real-world components, and analyse its theoretical behaviour.

The requirement is for a circuit that provides a minimum input resistance of 50k, a maximum gain of 0dB, and will use the common TL081 single op-amp. In all calculations we will use the worst-case values (Table 1) to give an upper bound on performance – typical performance will be better, but any differences will scale accordingly.

Characteristic	Symbol	Value	Units
Input Bias Current	I_{IB}	10	nA
Input Offset Voltage	V_{OS}	8	mV
Equivalent Input Noise Voltage (10Hz to 10kHz)	v_n	4	μV

Table 1: *Worst-case characteristics for Texas Instruments' TL081 single op-amp.*

When choosing components the nearest preferred values in the E24 series will be used.

Component	**Classic**	**New**
R_1	100k linear pot	51k
R_2	100k	10k linear pot
R_f	100k	51k

Table 2: *Components chosen for the two designs, using nearest preferred values.*

The results from calculation are presented in Table 3.

Characteristic	**Classic**	**New**
Input resistance range	50-100k	51-53.5k
Non-conformance	0.12	0.028
Offset due to I_{IB}	$556\mu V$	$83.6\mu V$
Noise	$8.03\mu V$	$5.84\mu V$
Noise Gain (max)	2 (6dB)	7.1 (17dB)
Offset due to V_{OS}	16mV	56.8mV

Table 3: *Performance of the classic and new input attenuator circuits. In most indicators the new circuit performs significantly better than the classic circuit. Noise Gain is about 12dB higher. Offset due to V_{OS} is higher too, although still insignificant for audio mixers, and trimmable for DC mixers.*

Clearly the new circuit is better than the old in almost all respects. The input resistance varies but by a much smaller amount – improvement can be achieved by reducing the pot value, increasing the value of the input resistor, or adding an input buffer. The conformance to the ideal is about five times better, and could also be improved by increasing R_1 or decreasing R_2. The DC offset is much less, about one-seventh, and could be improved by reducing the value of R_2. Thermal noise is reduced by almost 3dB due to the lower pot resistance. However, the Noise Gain is worse, by about 12dB, due again to the lower pot resistance. This can be minimised through the careful choice of op-amp.

5 Experimental Results

Theory is all very well, but a real circuit demonstrates how well the theory behaves in practice.

A single input realisation of the circuit shown in Figure 5 was constructed using readily-available components. One half of a TL082C dual op-amp was used, both R_f and R_{in} were 47k 1% resistors, and the pot was a linear 10k 16mm carbon track type, typical of the type used in low- to medium-cost modular synthesizers.

Measurements were taken for a DC input, with AC performance expected to be similar, subject to the usual issues surrounding finite bandwidth.

5.1 Results

5.1.1 DC Gain

With $V_{in} = 0V$ the output was measured as +2.14mV. This is primarily due to the Noise Gain applied to the op-amp's offset voltage.

With V_{in} set to 5.02V the full range of the output with the pot at the limits were measured as +1.83mV with the wiper at ground, and -4.98V with the wiper at the opposite end.

The two fixed resistors were carefully measured using a Solartron 7150 6-digit bench multimeter. Their values were:

$$\begin{aligned} R_f &= 47k00 \\ R_{in} &= 47k39 \end{aligned}$$

Given the actual values of the fixed resistors the theoretical circuit gain is

$$\begin{aligned} Gain &= -\frac{R_f}{R_{in}} \\ &= -\frac{47k00}{47k39} \\ &= -0.9918 \end{aligned}$$

With $V_{in} = 5.02$V we would expect a maximum output of

$$5.02V * -0.9918 = -4.979V$$

This is confirmed by the experimental result of -4.98V.

5.1.2 Input Resistance

The input resistance was measured at three key points of the pot's position (Table 4).

Rotation (%)	Input Resistance
0%	47k5
50%	50k1
100%	47k7

Table 4: *R_{in} for various pot positions.*

This agrees well with theory, which gives a minimum of 47k, and a maximum of 50k3.

6 Prior Art

During the writing of this paper it has been brought to my attention that the *reversed-pot* approach developed here has for many years been adopted by electric guitar manufacturers. In these circuits (*e.g.*, `http://www.gibson.com/Files/schematics/thinline.gif`) no op-amp is used, to save costs, and the reason for using this circuit is to avoid shorting the input to ground when one of the level pots is set to 0.

This approach has also been identified in the Mackie 1604VLZ mixer[5] in the aux-send pots in the channel strips. In this case the wiper resistors are 5k1 and the pots are 50k 'G'-law[6] so clearly some degree of law-bending is going on.

7 Acknowledgements

I would like to thank Tony Allgood, Tim Stinchcombe and Phil Macphail for their valuable time in discussing various aspects of this paper at Synth-DIY UK 2011, and to Harry Bissell through online correspondence on the synth-diy email list.

A Classic Circuit Design Equations

The input resistance can be written almost directly from the schematic in Figure 9.

[5]Schematics can be found at `http://www.gyraf.dk/schematics/schematics.html` in the *Mixers* section.

[6]The G taper is for graphic equalizers (hence the 'G'). It is symmetrical, with 50% resistance at 50% rotation, 10% resistance at 30% rotation, and 90% resistance at 70% rotation.

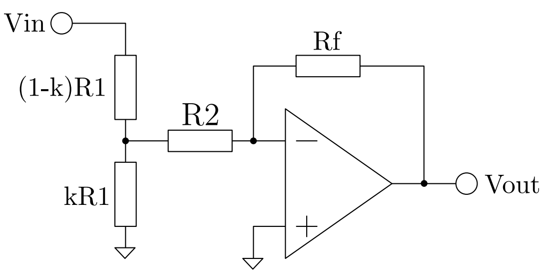

Figure 9: *Splitting R_1 into two helps understand the circuit analysis.*

$$R_{in} = (1 - k)R_1 + R_2 \parallel kR_1$$

Expanding out the parallel combinations, we get

$$R_{in} = \frac{R_1 R_2 + (k - k^2)R_1^2}{R_2 + kR_1} \tag{A.1}$$

The resistance seen by the *-ve* pin can also be seen from the schematic:

$$R_- = R_f \parallel (R_2 + kR_1 \parallel (1 - k)R_1)$$

Again, expanding out the parallel combinations and simplifying,

$$R_- = \frac{R_f R_2 + (k - k^2)R_f R_1}{R_2 + (k - k^2)R_1 + R_f} \tag{A.2}$$

The gain of the circuit, $\frac{V_{out}}{V_{in}}$, requires a little more effort. For the ideal op-amp we assume the *-ve* terminal is held at virtual earth. So,

$$V_{out} = -I_f R_f$$

Again, with an ideal op-amp no current flows into or out of the input pins, so

$$V_{out} = -I_2 R_f$$

where

$$I_2 = \frac{V_w}{R_2}$$

To determine I_2 we need to know voltage at the wiper, V_w. This is simply the output voltage of a potential divider:

$$V_w = V_{in} \frac{kR_1 \parallel R_2}{(kR_1) \parallel R_2 + (1 - k)R_1}$$

Putting these all together we arrive at

$$\frac{V_{out}}{V_{in}} = \frac{-kR_f}{R_2 + (k - k^2)R_1} \tag{A.3}$$

If we assume that $R_2 \gg R_1$ then we get close to the ideal behaviour:

$$\frac{V_{out}}{V_{in}} \approx \frac{-kR_f}{R_2}$$

B New Circuit Design Equations

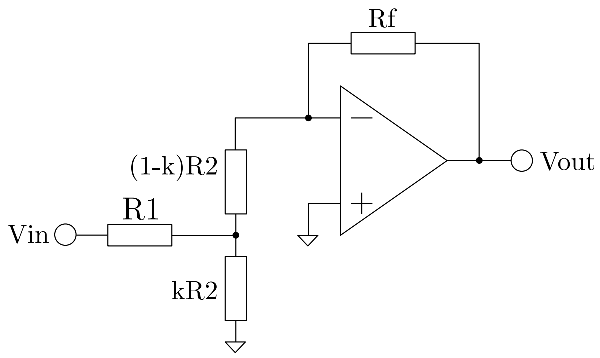

Figure 10: *Splitting R_2 into two helps understand the circuit analysis.*

The input reistance, R_{in}, is similar to the classic circuit,

$$R_{in} = R_1 + kR_1 \parallel (1 - k)R_2$$

which can be simplified to

$$R_{in} = R_1 + (k - k^2)R_2 \tag{B.1}$$

The expression for the resistance seen by the *-ve* pin is rather more complicated though:

$$R_- = \frac{(k - k^2)R_f R_2^2 - R_f R_1 R_2}{(k - k^2)R_2^2 - (R_1 + kR_f)R_2 - R_f R_1} \tag{B.2}$$

Finally, the transfer function happens to be of the same form as for the classic circuit, but note the exchange of R_1 and R_2:

$$\frac{V_{out}}{V_{in}} = \frac{-kR_f}{R_1 + (k - k^2)R_2} \tag{B.3}$$

Again, if we assume that $R_1 \gg R_1$ then we get close to the ideal behaviour:

$$\frac{V_{out}}{V_{in}} \approx \frac{-kR_f}{R_1}$$