Analog Circuits Handbook
模拟电路手册

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序言

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Preface

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Contents

Chapter 1 Measurement Methods

1.1 General Measurement Methods for Impedance

Relevant resources:

(1) Digilent Reference: Using the Impedance Analyzer (https://digilent.com/reference/test-and-measurement/guides/waveforms-impedance-analyzer)

(2) 3 Ways to Measure Inductance (https://www.wikihow.com/Measure-Inductance)

(3) Inductance Measurement Using an LCR Meter and a Current Transformer Interface (https://www.nist.gov/system/files/documents/calibrations/WaltripProceedings.pdf)

1.2 Inductor DCR (Direct Current Resistance) Measurement

DCR, standing for direct current resistance, just as its name implies, is the resistance of the inductor when a DC current flows through it. It is different from the equivalent series resistance (ESR) measured in the last section, which is the real part of the impedance of the inductor for AC signals. Concretely speaking, we use DCR to calculate the DC bias (large-signal) of a certain circuit, but use ESR to calculate the small-signal response.

1.3 High-precision Inductance Measurement

High-precision Measurement circuit is shown in Fig.1.1 and the inductor DCR (or ESR) does not affect the measurement result.

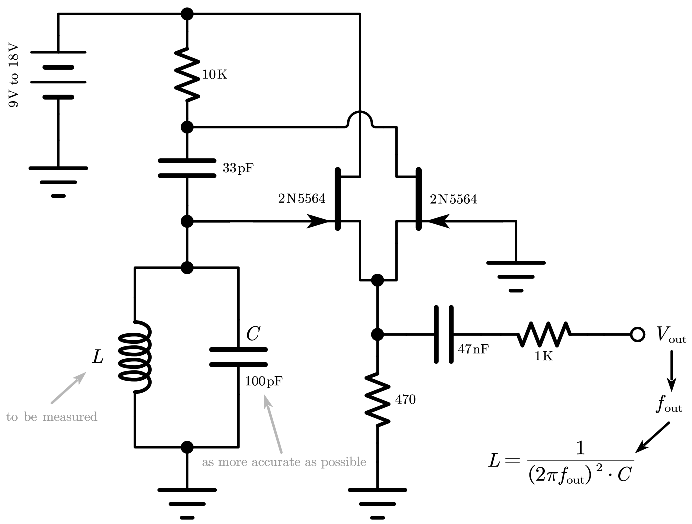


Figure 1.1: High-precision inductance measurement circuit

Below is an output frequency reference with $C = 100\text{pF}$ for different inductance range. Simply change the value of C to meet your requirements. For a precise measurement, high-precision capacitor is recommended.

Table 1.1: Output Frequency Reference for Different Inductance Range

L	1 nH	1 uH	1 mH	1 H	1000 H
f_{out}	503.29 MHz	15.915 MHz	503.29 KHz	15.915 KHz	503.29 Hz

1.4 Basic Current Sensing Circuit

Relevant resources:

(1) Current Sensing Circuit Concepts and Fundamentals (https://ww1.microchip.com/downloads/en/AppNotes/01332B.pdf)

(2) An Engineer's Guide to Current Sensing (https://www.ti.com/lit/eb/slyy154b/slyy154b.pdf?ts=1735114071672&ref_url=https%253A%252F%252Fcn.bing.com%252F)

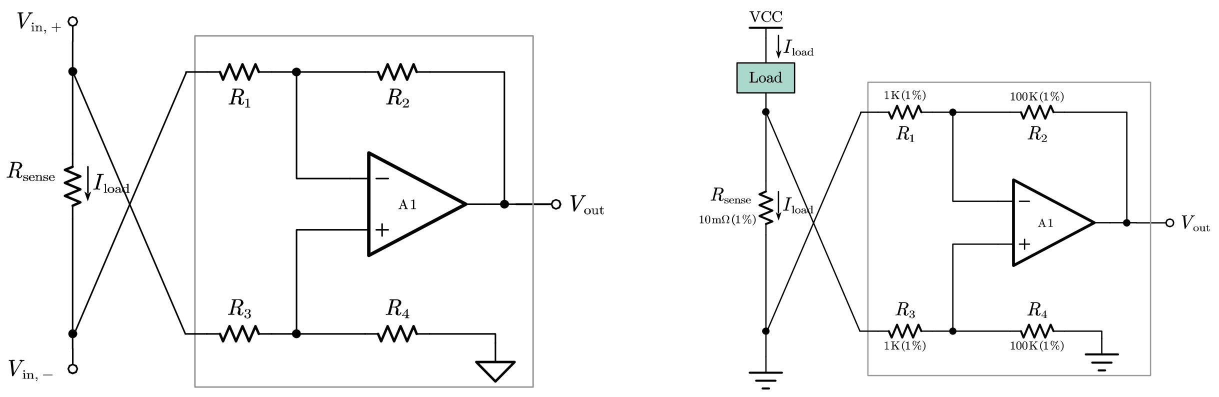
(3) Switch Mode Power Supply Current Sensing - Part 3: Current Sensing Methods (https://www.analog.com/media/en/technical-documentation/tech-articles/A59723-Part3-Switch-Mode-Power-Supply-Current-Sensing-Part-3-Current-Sensing-Methods.pdf)

(4) Understanding Current Sensing Applications & How to Choose the Right Device (https://www.ti.com/lit/ml/slap178/slap178.pdf?ts=1735143793758&ref_url=https%253A%252F%252Fcn.bing.com%252F)

(5) Current Sensing Techniques: A Review (https://ieeexplore.ieee.org/document/4797906)

1.4.1 General Considerations

Depicted in 1.2, we use a difference amplifier to sample the current. A general operational amplifier can be fine, but a precision one is recommended for better performance. We will see the error difference between the two in the following subsection.



(a) Simple current sensing schematic **(b) Typical application (low-side current sensing)**

Figure 1.2: One-opamp current sensing circuit

To gain a better ouput performance at low load current, ± VCC is recommended for the power supply of the amplifiers. Remark that $R_1 = R_3$ and $R_2 = R_4$ are required for the difference amplifier. In this case, the output voltage is given by:

$$V_{\text{out}} = \frac{R_2}{R_1} \cdot I_{\text{load}} R_{\text{sense}} \Longrightarrow I_{\text{load}} = \frac{V_{\text{out}}}{R_{\text{sense}} \cdot \frac{R_2}{R_1}} \tag{1.1}$$

setting $R_1 = R_3 = 1\ \text{K}\Omega$, $R_2 = R_4 = 100\ \text{K}\Omega$ and $R_{\text{sense}} = 0.01\ \Omega$ yileds:

$$I_{\text{load}} = V_{\text{out}} \tag{1.2}$$

where the unit of I_{load} is ampere (A) and the unit of V_{out} is volt (V).

1.4.2 LTspice Simulation

Now we consider two different operational amplifiers, NE5532 and OP07, to run the simulation in LTspice, where NE5532 is a general operational amplifier and OP07 is a precision one. Here are the specific details of their parameters:

Table 1.2: Parameters of OP07 and NE5532

OPA	V_{io}	I_{io}	I_{b}	A_v	Slew Rate
OP07	60 uV	0.8 nA	1.8 nA	400 V/mV	0.3 V/us
NE5532	0.5 mV	10 nA	200 nA	200 V/mV	9 V/us

(1) V_{io}: input offset voltage;
(2) I_{io}: input offset current;
(3) I_{b}: bias current;
(4) A_v: large-signal open-loop gain;
(5) Slew Rate: the maximum rate of output voltage change.

Construct the simulation circuit in LTspice as shown in Fig.1.3 and run the DC sweep. We can see the output and error difference between the two operational amplifiers in Fig.1.4, where the relative error η is defined as:

$$\eta = \frac{V_{\text{out}} - I_{\text{load}}}{I_{\text{load}}} \times 100\,\% \tag{1.3}$$

Since OP07 has a low input offset and bias, the measurement error is negligible. By contrast, the error with NE5532 is a bit obvious. This difference becomes more significant with $I \in [0,\ 0.1A]$, shown in Fig.1.5.

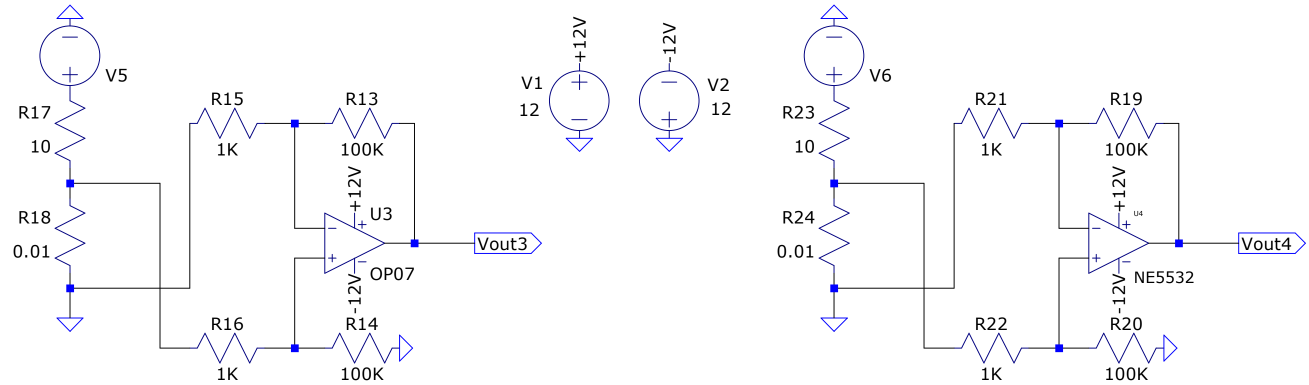
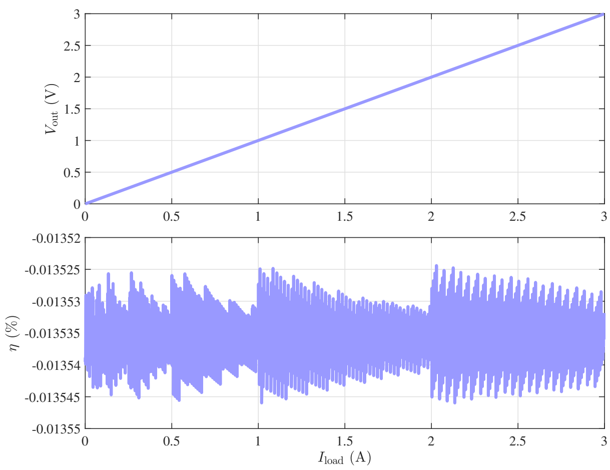
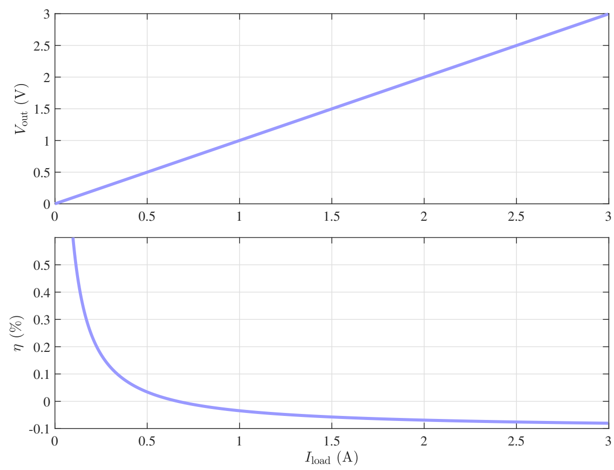


Figure 1.3: LTspice simulation of simple current sensing circuit

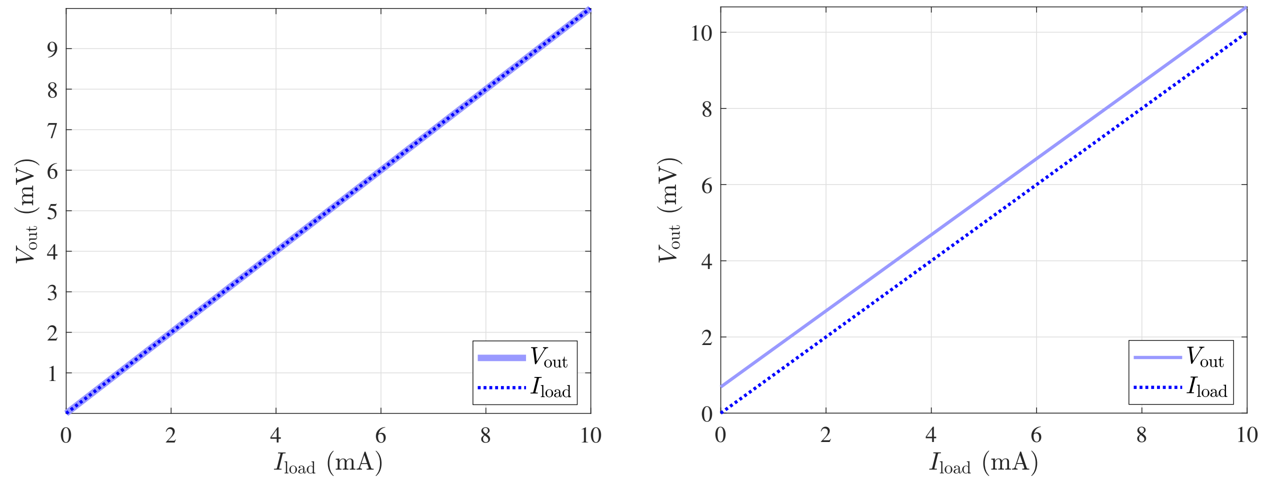


(a) Simulation results for OP07



(b) Simulation results for NE5532

Figure 1.4: Simulated ouput voltage and reletive error for $I_{\text{load}} \in [0\text{A},\ 3\text{A}]$



(a) Simulation results for OP07 (b) Simulation results for NE5532

Figure 1.5: Simulated ouput voltage and reletive error for $I_{\text{load}} \in [0\text{mA},\ 10\text{mA}]$

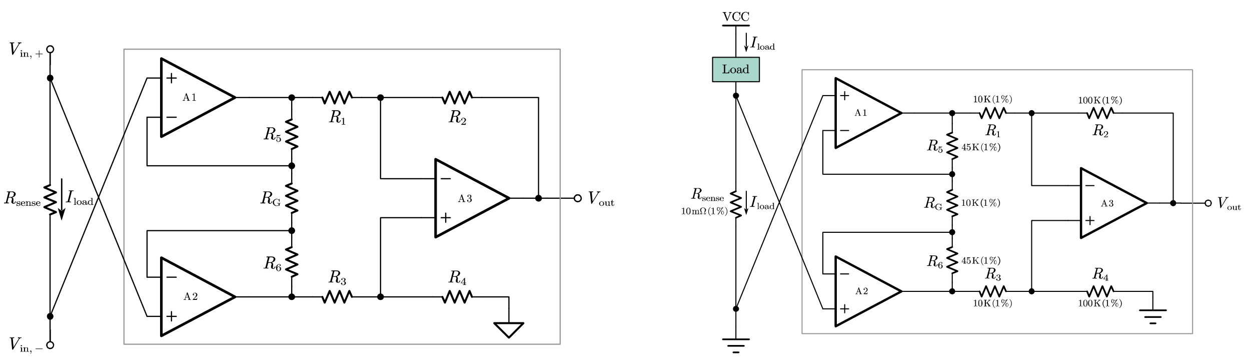
1.4.3 Actual Circuit Test

1.5 Precision Current Sensing Circuit

> **Relevant resources:** same as in the previous section (Sec.1.4 Simple Current Sensing Circuit)

1.5.1 General Considerations

Depicted in Fig.1.6, we use an instrumentation amplifier comprised of three amplifiers to precisely sample the current of a load. $R_1 = R_3$, $R_2 = R_4$, $R_5 = R_6$ are necessary, and $R_G \in [0.2\ \text{K}\Omega,\ 2\ \text{K}\Omega]$, $R_{\text{sense}} \in [1\text{m}\Omega,\ 1\ \Omega]$ are recommended for better performance.



(a) Circuit schematic (b) Typical application (low-side current sensing)

Figure 1.6: Precision current sensing circuit

The output voltage is given by:

$$V_{\text{out}} = \left(1 + \frac{2R_5}{R_G}\right)\frac{R_2}{R_1} \cdot I_{\text{load}} R_{\text{sense}} \Longrightarrow I_{\text{load}} = \frac{R_1}{R_2 R_{\text{sense}} \cdot \left(1 + \frac{2R_2}{R_1}\right)} \cdot V_{\text{out}} \tag{1.4}$$

1.5.2 LTspice Simulation

1.5.3 Actual Circuit Test

Chapter 2 Capacitors

2.1 Capacitance Multiplier

Relevant resources:

(1) A Review of Capacitance Multiplication Techniques (https://ieeexplore.ieee.org/document/8678969)

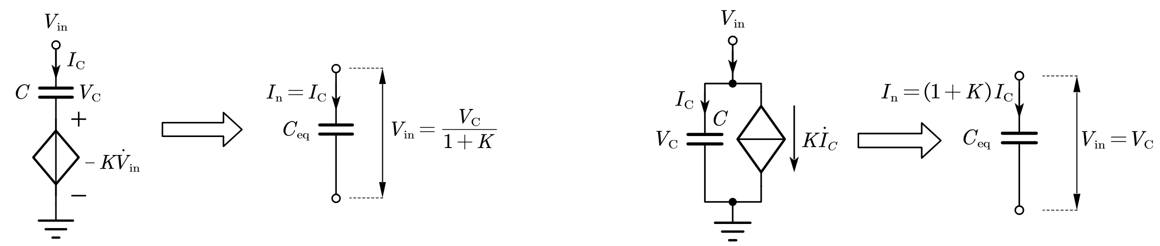
(2) Capacitance Multiplier with Large Multiplication Factor and High Accuracy (https://www.jstage.jst.go.jp/article/elex/15/3/15_15.20171191/_article)

(3) Active Capacitor Multiplier in Miller-compensated Circuits (https://ieeexplore.ieee.org/document/818917)

(4) The Capacitance Multiplier (https://audioxpress.com/article/the-capacitance-multiplier)

2.1.1 Basic Circuits and Principles

Below are two basic concepts for capacitance multiplication:



(a) Voltage Mode **(b) Current Mode**

Figure 2.1: Basic Two Concepts for Capacitance Multiplier

Thus, we obtain the equivalent capacitance as:

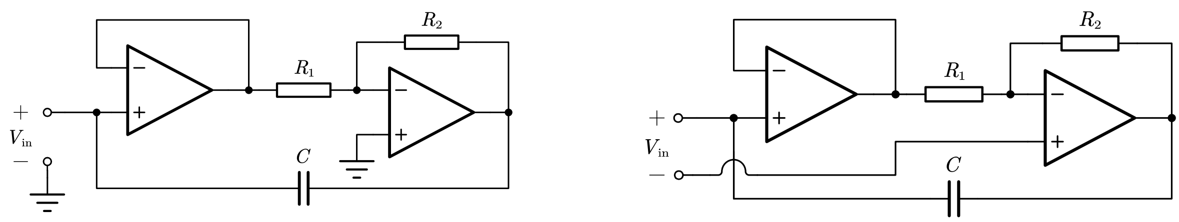
$$\text{voltage mode:} \quad C_{eq} = \frac{I_n}{SV_n} = \frac{I_C}{S\frac{V_C}{1+K}} = (1+K)\,C, \quad K > 0 \tag{2.1}$$

$$\text{current mode:} \quad C_{eq} = \frac{I_n}{SV_n} = \frac{(1+K)I_C}{SV_C} = (1+K)\,C, \quad K > 0 \tag{2.2}$$

A simple implementation of cap multiplier, depicted in Fig.2.2 (a), combining a unit-gain buffer (voltage fllower) and an inverting amplifier, uses a voltage mode. yielding the equivalent capacitance:

$$A = -\frac{R_2}{R_1} = -K \Longrightarrow C_{eq} = \left(1 + \frac{R_2}{R_1}\right) C \tag{2.3}$$

where $A = -\frac{R_2}{R_1}$ is the closed-loop gain of the inverting amplifier. Since inverting amplifier has a low input impedance, the unit-gain buffer is a necessary. To change it into a two-terminal element, just replace GND with the negtive terminal of the input voltage, e.g. $V_{in,-}$, as shown in Fig.2.2 (b).



(a) One-terminal Cap Multiplier **(b) Two-terminal Cap Multiplier**

Figure 2.2: A Simple Implementation of Capacitance Multiplier

2.1.2 Multisim Simulation

Considering Figure 2.2 (b), set the parameters as Table 2.1. Then we connect it to the RC series circuit to perform an AC sweep to test the capacitance value. The Simulation Circuit is shown in Figure 2.3.

Table 2.1: Simulation Parameters of Capacitance Multiplier

C	R_1	R_2	Operation Amplifier
10 nF	1 KΩ	11 KΩ	LM258P

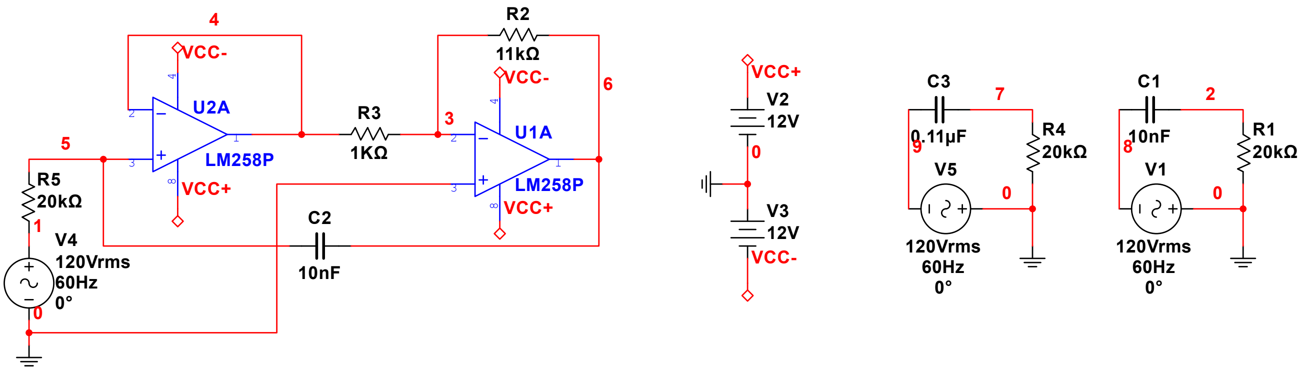


Figure 2.3: Simulation Circuit of Cap Multiplier

Export the simulation data and plot the frequency response (bode plot) of the series RC circuit, as shown in Figure 2.4. The theoretical value of the capacitance of the cap multiplier is 110 nF, and the simulation result confirmed this point.

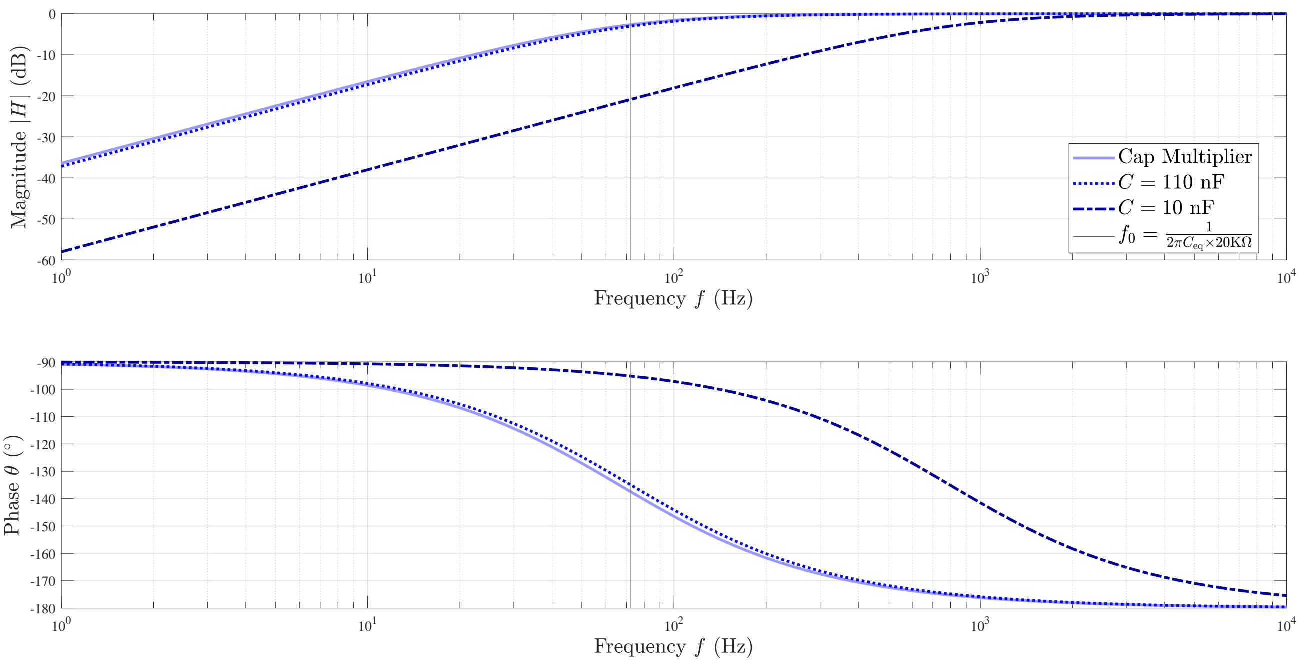


Figure 2.4: AC Sweep of the Cap Multiplier

Chapter 3 Inductors

3.1 Gyrator-Based Inductor

Relevant resources:

(1) Class AB Gyrator-Based Active Inductor (https://doi.org/10.1109/INMMIC.2015.7330365)

(2) Gyrator Based Inductor (https://www.academia.edu/4246523/gyrator_based_inductor)

Chapter 4 Oscillators

4.1 The Wien Bridge Oscillator

You must have seen that a number of resistors and capacitors can be connected together with an inverting amplifier to produce an oscillating circuit. Wien bridge oscillator is one of the simplest sine wave oscillators which uses an RC network in place of the conventional LC tuned tank circuit to produce a sinusoidal output waveform.

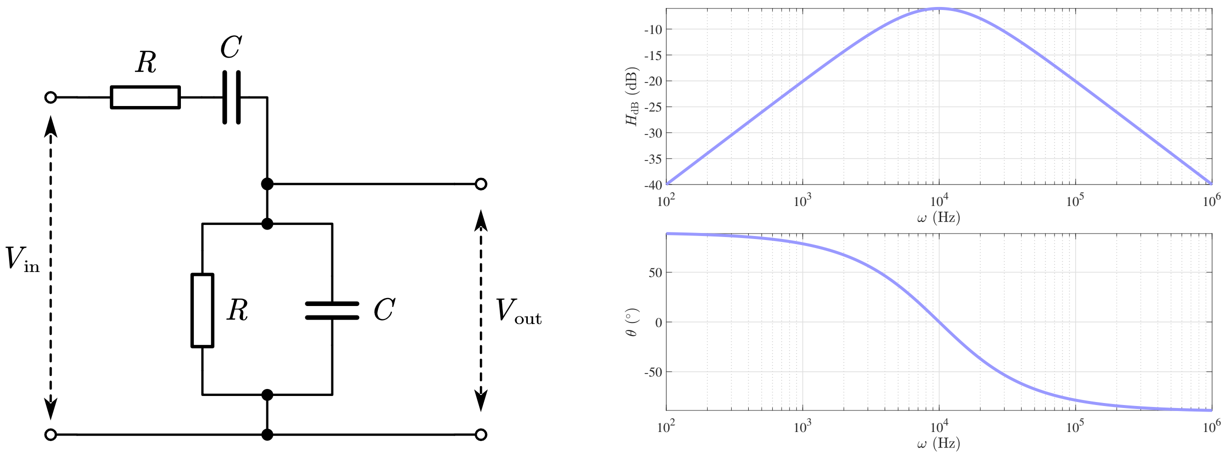
The Wien Bridge Oscillator is based on a noninverting amplifier, using a series RC circuit connected with a parallel RC of the same component values as a feedback circuit.

4.1.1 Basic Circuit and Principles

Consider the RC circuit in Figure 4.1 (a), the voltage gain H of the series RC circuit is:

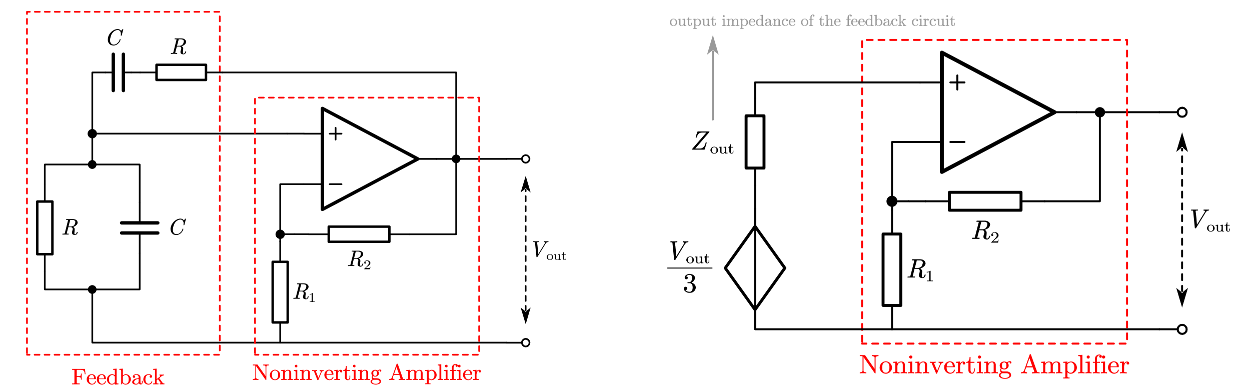
$$H(j\omega) = \frac{R \parallel (\frac{1}{j\omega C})}{R + \frac{1}{j\omega C} + R \parallel (\frac{1}{j\omega C})} = \frac{1}{1 + \frac{(R^2 - \frac{1}{\omega^2 C^2}) + \frac{2R}{j\omega C}}{\frac{R}{j\omega C}}}, \quad H|_{\omega = \frac{1}{RC}} = \frac{1}{1 + \frac{\frac{2R}{j\omega C}}{\frac{R}{j\omega C}}} = \frac{1}{3} \tag{4.1}$$

Defined to obtain a 0° phase shift, the resonant frequency f_0 is the key to Wien bridge oscillator. And at the point we have $H = \frac{1}{3}$. Let's set $R = 10\ \text{K}\Omega$, $C = 10\ \text{nF}$ and sketch the bode plot of this RC circuit in Figure 4.1 (b).



(a) RC Feedback Circuit **(b) Bode Plot of the RC Circuit**

Figure 4.1: Wien Bridge Oscillator's Feedback Circuit



(a) Wien Bridge Oscillator **(b) Equivalent Circuit of Wien Bridge Oscillator**

Figure 4.2: Wien Bridge Oscillator and Its Equivalent Circuit

Since a noninverting amplifier has extreme high input impedance and low output impedance, the coupling effect of the two circuits is negligible. In other words, the output impedance of noninverting amplifier (combing with the input impedance of feedback circuit), and the output impedance of feedback circuit (combing with input impedance of noninverting amplifier) can be ignored. Thus, the Wien bridge oscillator, depicted in Figure 4.2 (a), has the equivalent circuit shown in Figure 4.2 (b).

The oscillation frequency f_0 of the Wien Bridge Oscillator is given by:

$$\omega_0 = \frac{1}{RC}, \quad f_0 = \frac{1}{2\pi RC} \tag{4.2}$$

As the voltage gain of noninverting amplifier is:

$$A_v = 1 + \frac{R_2}{R_1} \tag{4.3}$$

yielding **the start-oscillation condition**:

$$A_v > 3 \Longleftrightarrow R_2 > 2R_1 \tag{4.4}$$

Assuming R_2 is slightly greater than $2R_1$, and there is a noise signal consists of a series of frequency, including $f_0 = \frac{1}{2\pi RC}$. Then at the selected resonant frequency f_0, $\frac{1}{3}A_v > 1$, so the positive feedback will cancel out the negative feedback signal, causing the circuit to oscillate, until it reaches a voltage saturation (dependent on power supply). However, at the other frequency, $\frac{1}{3}A_v < 1$ so the negative feedback will cancel out the positive, resulting other frequency signal fading away.

The closer the ratio $\frac{R_2}{R_1}$ is to 2^+, the better the waveform, but the longer the start-up time. Define $a = \frac{1}{3}\left(1 + \frac{R_2}{R_1}\right)$ as the periodic gain, a not-bad approximation for the start-up time is:

$$t_{\text{start}} = \frac{1}{f} \log_{a^3}\left(\frac{V_{\text{limit}}}{V_{\text{noise}}}\right) - 0.02 \tag{4.5}$$

where the unit of t is seconds, V_{limit} is the limit amplitude of output voltage, V_{noise} is the amplitude of noise.

By the way, if R_2 exceeds $2R_1$ too much, for example $R_2 = 3R_1$, the output waveform will be seriously distorted. Also, due to the slew rate limitations of operational amplifiers, frequencies above 1 MHz are unachievable without the use of special high frequency op-amps.

4.1.2 Multisim Simulation of the Basic Circuit

Set the parameters in Figure 4.2 (a) as below, and run the simulation.

Table 4.1: Simulation Parameters of Wien Bridge Oscillator

R	C	R_1	R_2	Operation Amplifier	VCC
10 KΩ	10 nF	10 KΩ	20.1 KΩ	LM258P	± 12 V

The start-up time is about 570 ms, shown in Figure 4.3.

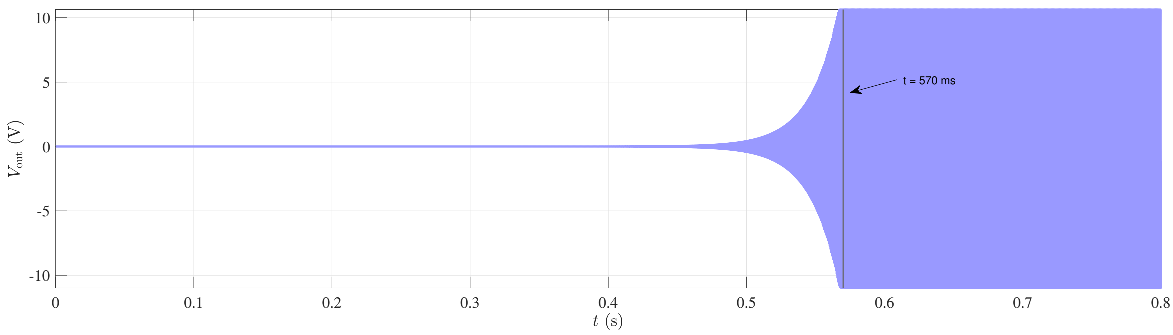


Figure 4.3: Start-up Time of Wien Bridge Oscillator

Export the simulation data, and perform a spectrum and distortion analysis in Matlab. Then we obtain the waveform and spectrum shown in Figure 4.4.

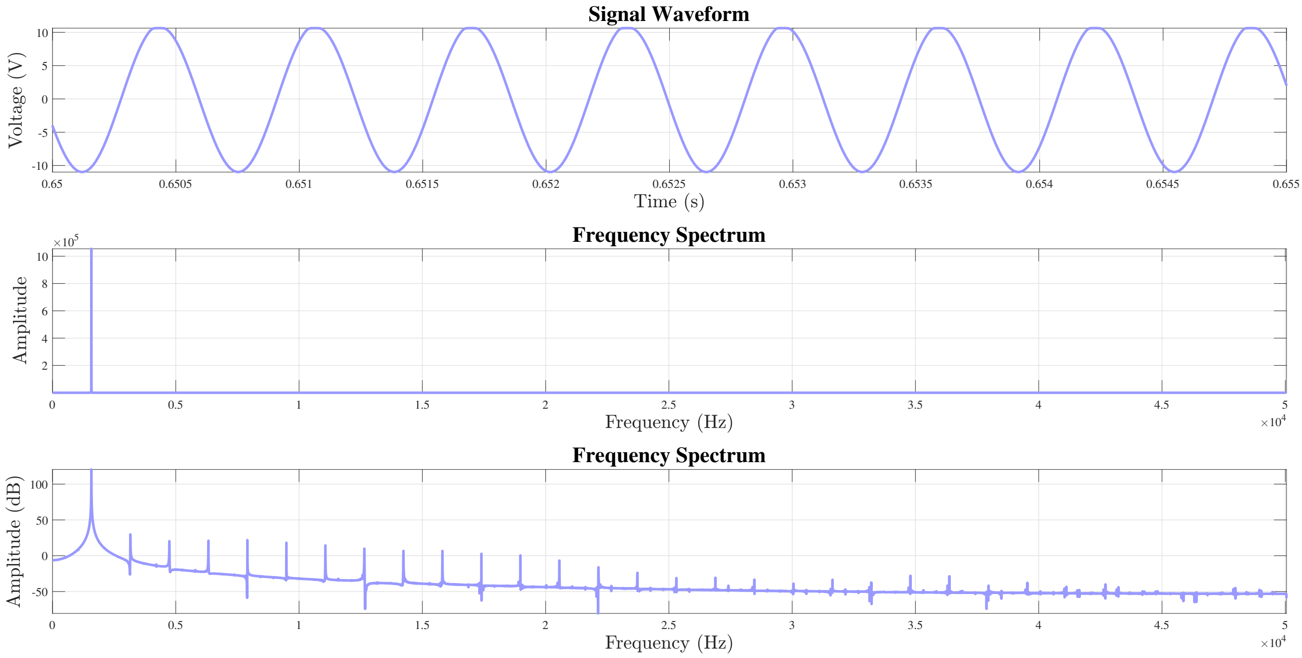


Figure 4.4: Spectrum Analysis of the Simulation Circuit

As we can see, the main output waveform is a sine wave at the resonant frequency f_0, the simulated oscillation frequency is:

$$\begin{aligned} f_{\text{simu}} &= 1.5758 \text{ KHz} \\ f_{\text{theo}} &= 1.5915 \text{ KHz} \\ \eta &= \frac{f_{\text{simu}} - f_{\text{theo}}}{f_{\text{theo}}} = -0.98 \ \% \end{aligned} \tag{4.6}$$

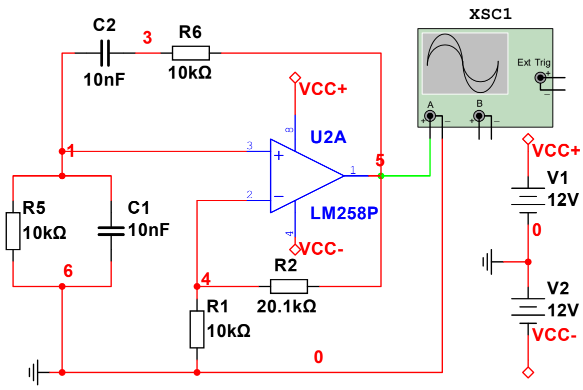
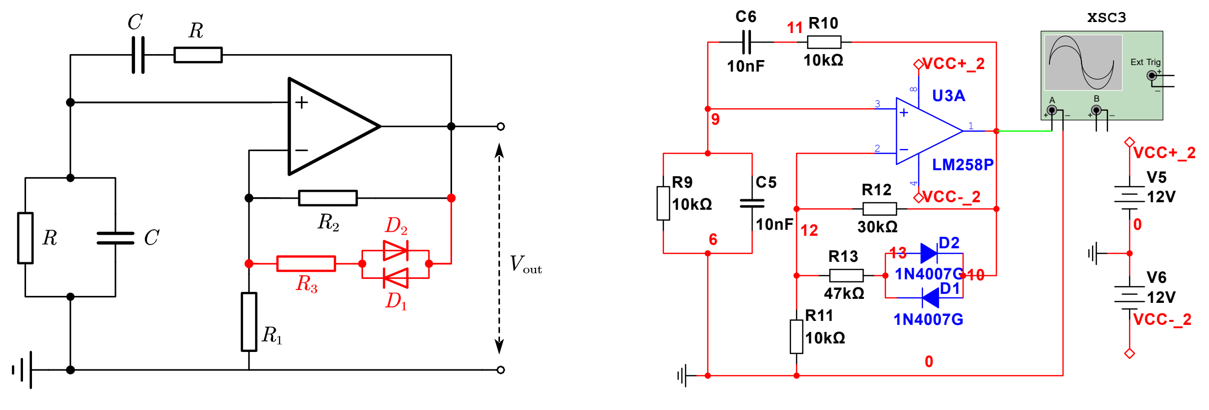


Figure 4.5: Simulation Circuit of Wien Bridge Oscillator

4.1.3 Decrease the Start-up Time

We have noted that the output waveform is distorted when R_2 exceeds $2R_1$ too much, but the start-up time is too long when R_2 is too closed to $2R_1$. Therefore, we need to optimize the circuit to achieve a better waveform and shorter start-up time, exemplified in Figure 4.6.

In Figure 4.6 (a), we added a resistance R_3 and two diodes D_1 and D_2. When the output voltage amplitude is less than the threshold voltage of diodes V_{D}, the diodes are off, and the circuit is the same as the basic circuit. When the output is greater than V_{D}, the diodes are on, and the resistance of R_3 is added to the circuit (parallel with R_2), which reduces the gain of amplifier.



(a) Optimized Circuit

(b) Simulation Circuit

Figure 4.6: Optimize the Start-up Time of Wien Bridge Oscillator

Simulation circuit is shown in Figure 4.6 (b), the start-up time is reduced to about 10 ms (see Figure 4.7), and the output waveform is shown in Figure 4.8.

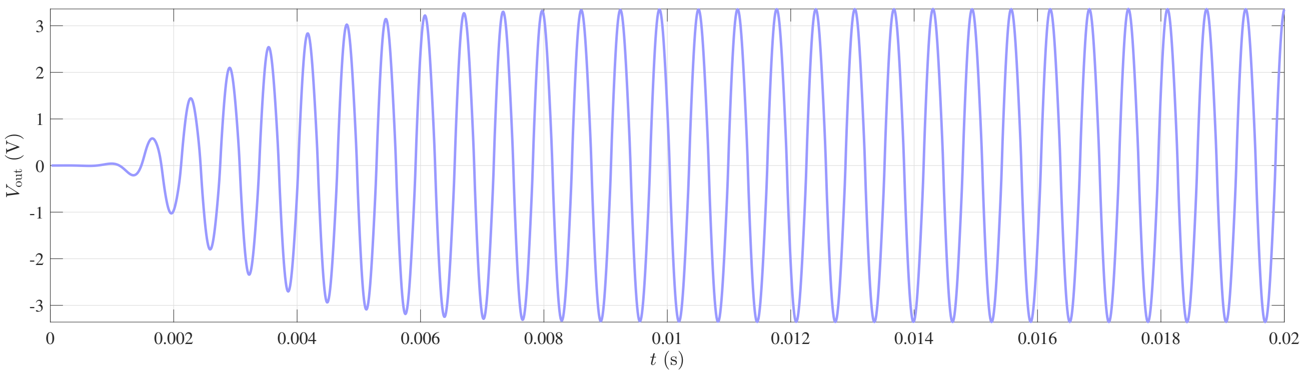


Figure 4.7: Optimized Start-up Time

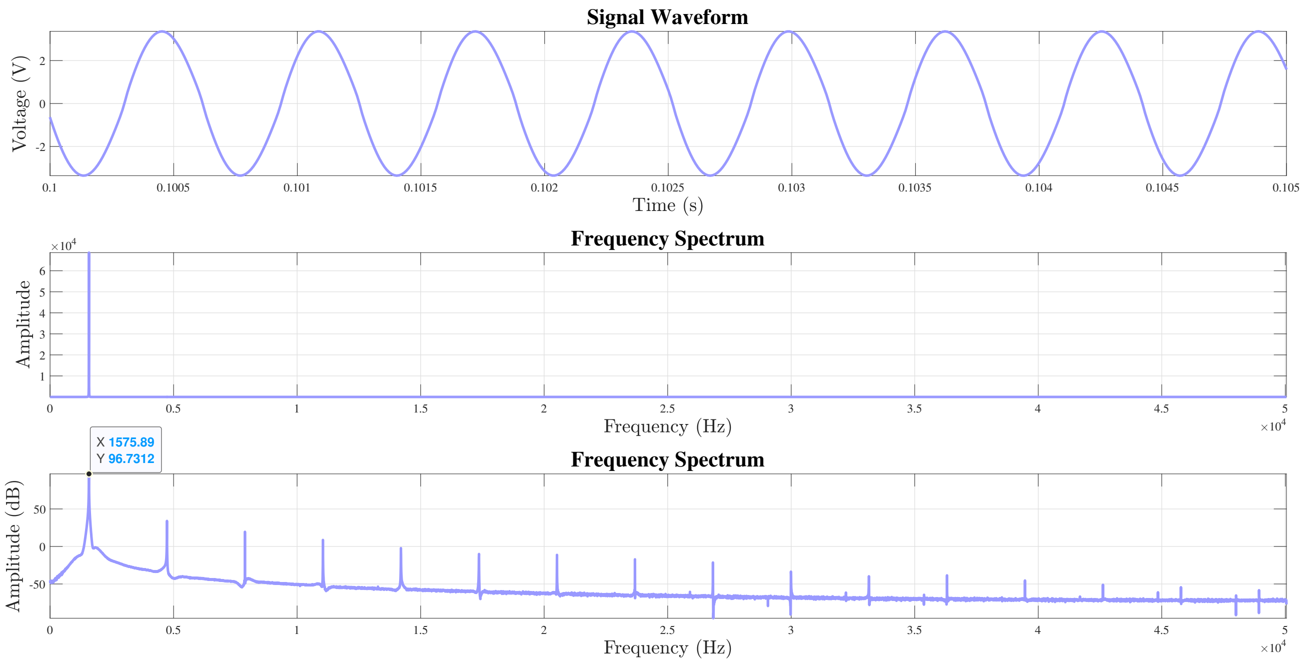


Figure 4.8: Spectrum Analysis of Optimized Circuit

Although the output frequency still focuses on f_0 and the distortion completely disappears, we have to note that the amplitude of the output waveform is significantly reduced. If larger amplitudes are desired, a resistor can be added to divide V_{out} to a suitable level, leading to a larger output amplitude (see reference [?] page 661). For more alternative methods to optimize the start-up time, see https://blog.csdn.net/qq_29356039/article/details/132611987.

4.1.4 Generate a Square Wave

An R_2 greater than $2R_1$ will result in the output waveform being clipped at the output voltage limitations. In other words, if we let $R_2 \gg 2R_1$, the waveform becomes a square wave. To prove our surmise, reset $R_1 = 1$ KΩ, $R_2 = 30$ KΩ in Table 4.1, without changing the other parameters. We obtain the output waveform shown in Figure 4.9.

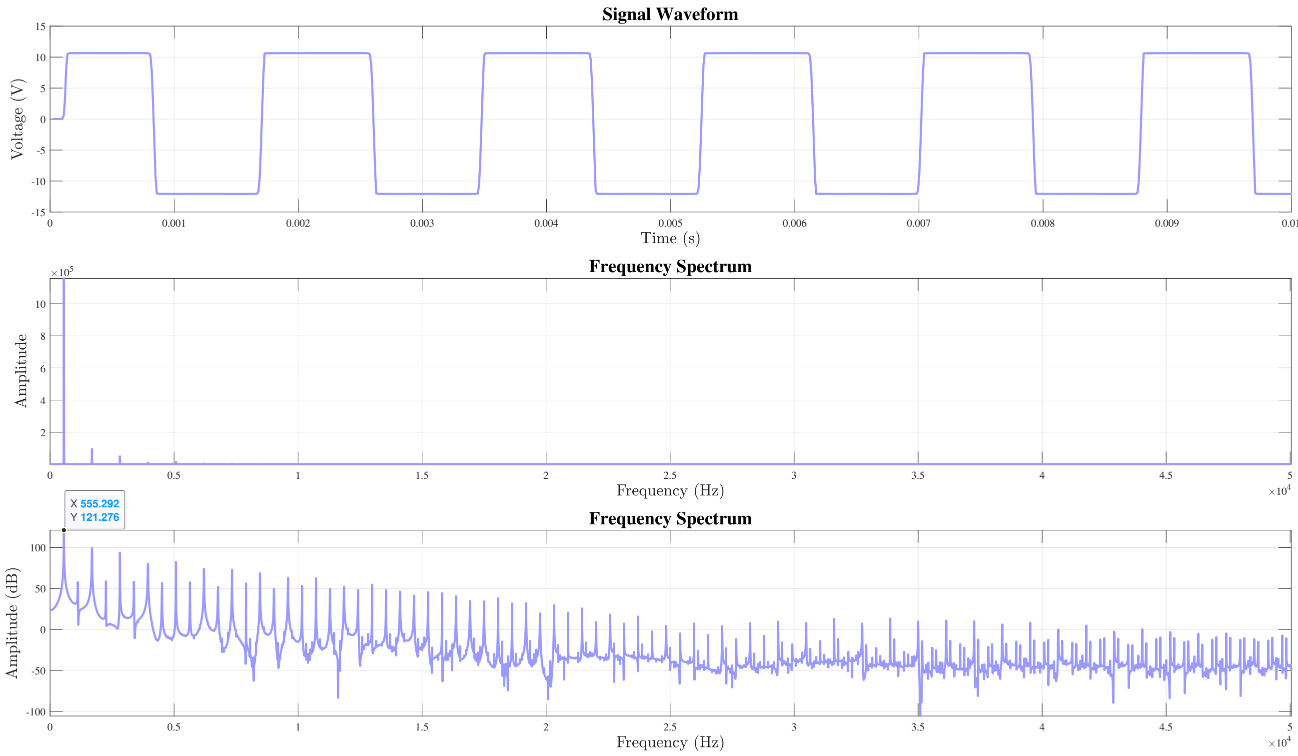


Figure 4.9: Square Wave Output of Wien Bridge Oscillator

Since the frequency of the square wave is difficult to calculate and control, the Wien Bridge Oscillator is not suitable for generating square waves in practical applications. To obtain an output signal with DC offset, see https://blog.csdn.net/qq_29356039/article/details/132611987 for more details.

Chapter 5 Current Source

5.1 Howland Current Pump ($0 \sim 1$mA)

Relevant resources:

(1) The Howland Current Pump (https://www.allaboutcircuits.com/technical-articles/the-howland-current-pump/)

(2) Precision Current Sources and Sinks Using VoltageReference (https://www.ti.com/lit/an/snoaa46/snoaa46.pdf?ts=1735325810030&ref_url=https%253A%252F%252Fwww.google.com%252F)

(3) Implementation and Applications of Current Sources and Current Receivers (https://www.ti.com/lit/an/sboa046/sboa046.pdf)

5.2 The Jim Williams Current Source ($0 \sim 10$mA)

5.3 Voltage Controlled Current Sink (0A ∼ 1A, up to 10A)

Relevant resources:

(1) Implementation and Applications of Current Sources and Current Receivers (Page 8)
(https://www.ti.com/lit/an/sboa046/sboa046.pdf)

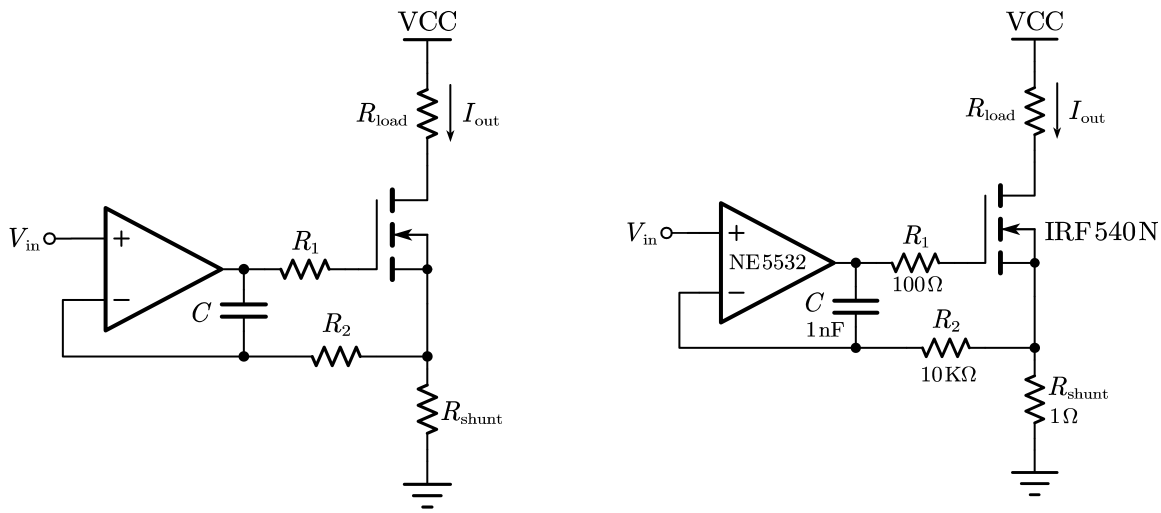
(2) Designing a Voltage Controlled Current Source Circuit
(https://www.circuits-diy.com/designing-a-voltage-controlled-current-source-circuit/)

(3) Design a Voltage Controlled Current Source Circuit using Op-Amp
(https://pic-microcontroller.com/design-a-voltage-controlled-current-source-circuit-using-op-amp/)

5.3.1 Circuit Schematic

The current source schematic is shown in Fig.5.1 (a) and the output current is given by:

$$I_{\text{out}} = \frac{V_{\text{in}}}{R_{\text{shunt}}}, \quad \forall\ I_{\text{out}}(R_{\text{load}} + R_{\text{shunt}}) \leq V_{\text{CC}} \tag{5.1}$$



(a) Circuit schematic **(b) Typical application**

Figure 5.1: General voltage controlled current source (1A-level current sink)

For the typical application shown in Fig.5.1 (b), $R_{\text{shunt}} = 1\ \Omega$, yielding:

$$I_{\text{out}} = \frac{V_{\text{in}}}{1\ \Omega} = V_{\text{in}} \tag{5.2}$$

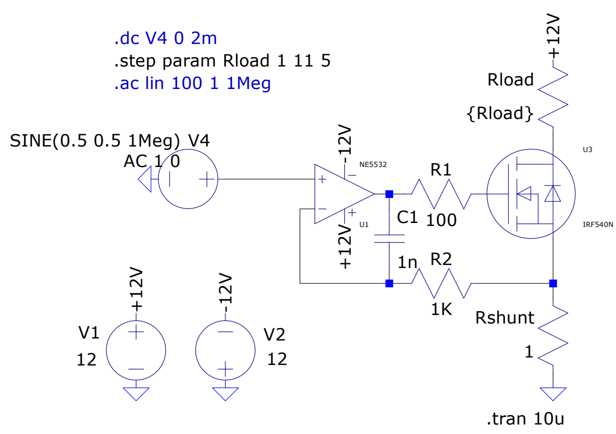
where the unit of V_{in} is in volts (V) and the unit of I_{out} is in amperes (A). Note that single power supply is completely for the opamp to ensure the proper operation in the low voltage range (we will see this in the following contents).

Although this current source is widely used for 1A-level current output, through reducing R_{shunt} and optimizing the on-resistance $R_{\text{DS, on}}$ of the MOSFET, e.g., IRF3205PBF ($R_{\text{DS, on}} = 8\ \text{m}\Omega$) or IRF1407PBF ($R_{\text{DS, on}} = 7.8\ \text{m}\Omega$), it is capable of outputting a current in 10A or larger.

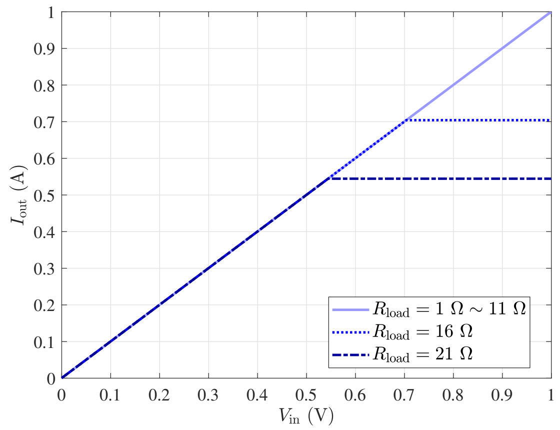
Also, to obtain better performance in the high-frequency range and low voltage range, we can use a precision opamp with high slew rate and low input offset, such as LT1102 (https://www.analog.com/en/products/lt1102.html) and LT1115 (https://www.analog.com/en/products/lt1115.html).

5.3.2 LTspice Simulation

The simulation circuit and its output Characteristics are shown in Fig.5.2. Perform AC sweep of the input voltage V_{in} from 0V to 1V, and sketch the network function $H = \frac{I_{\text{out}}}{V_{\text{in}}}$. The results are shown in Fig.5.3.



(a) Simulation Schematic



(b) Operation Characteristics

Figure 5.2: LTspice simulation circuit and results

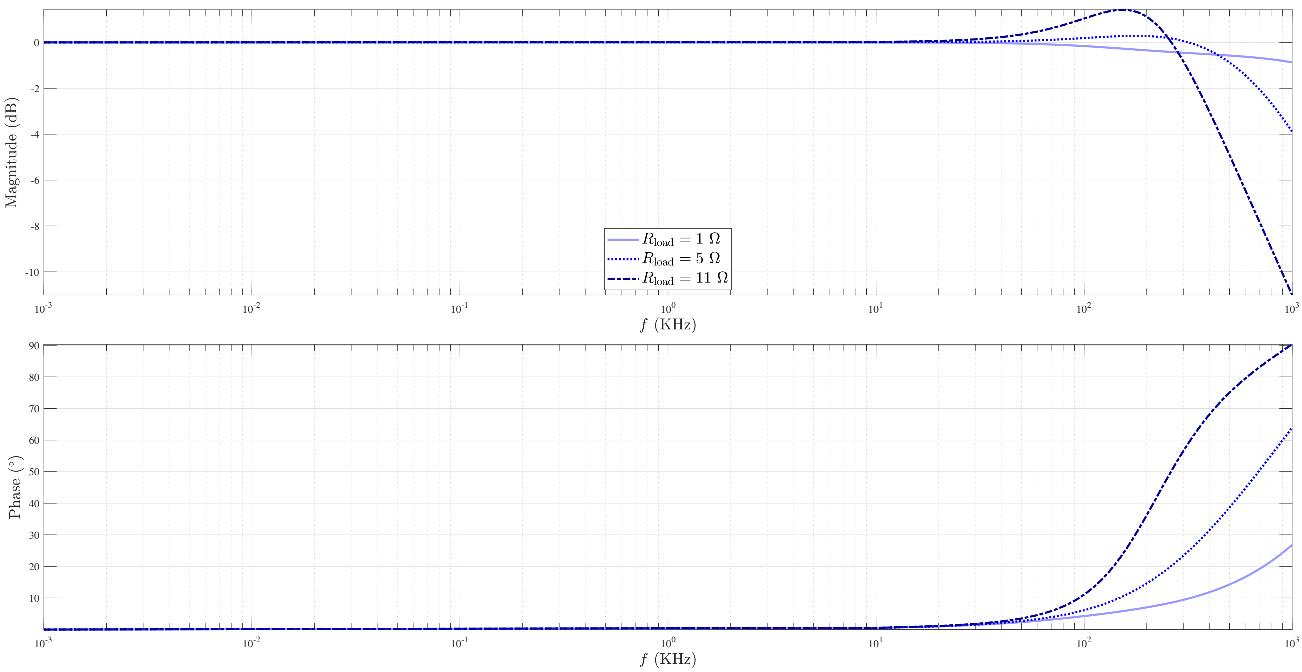


Figure 5.3: Output amptitude along with the input frequency for different load resistances

In addition, the capacitor C_1 is added to the circuit to stabilize the output current and limit the amptitude. The transient analysis results with and without the capacitor are shown below:

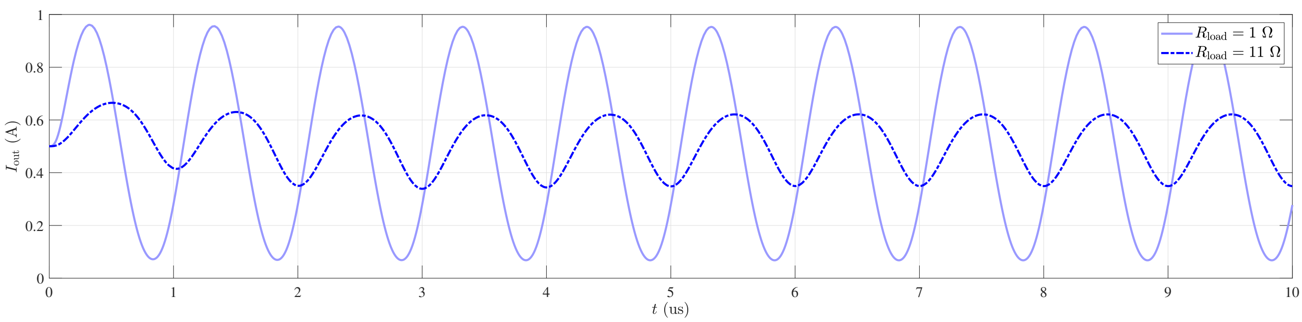


Figure 5.4: Transient response with $C = 1$ nF for 1MHz sine wave input, $V_{\text{in}} \in [0,\ 1\text{ V}]$

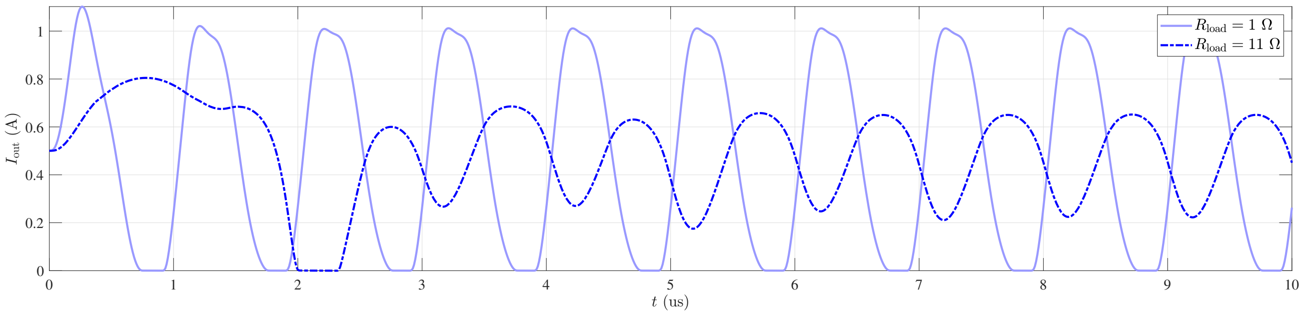


Figure 5.5: Transient response without C for 1MHz sine wave input, $V_{\text{in}} \in [0,\ 1\ \text{V}]$

Apparently, the output current becomes more stable and low-distortion if we add $C = 1$ nF.

What's more, since the output voltage of the opamp is between 0 V and 9 V (see Fig.5.6) for $I_{\text{load}} \in [0,\ 5\ \text{A}]$, a single power supply is completely enough for the common usage.

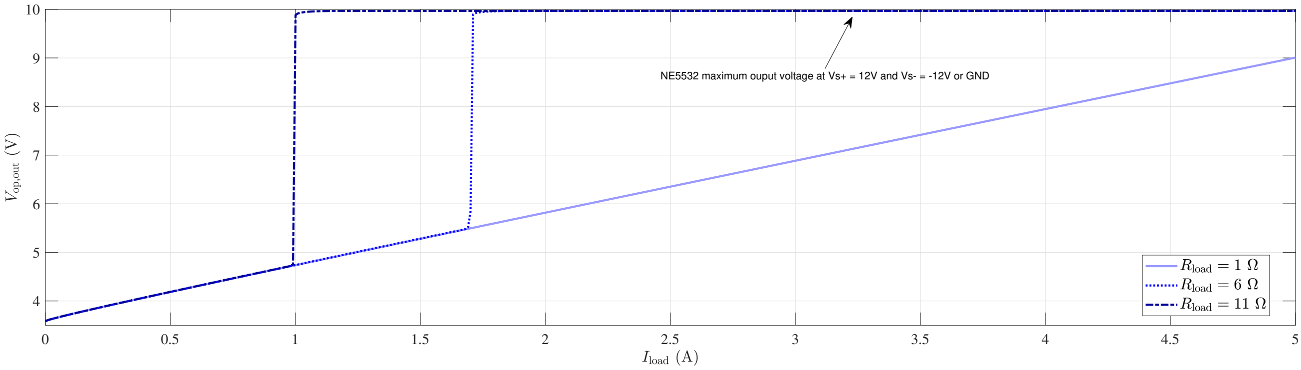


Figure 5.6: The output voltage of NE5532 as a function of I_{load} with V_{S+} =12V and $V_{S-} = -12V$ or GND

5.3.3 Perforated Board Verification

The perforated board layout is shown in Fig.5.1. The output current is measured by a multimeter and the results are shown in Fig.5.1 (c). The output current is $I_{\text{out}} = 0.99$ A, which is consistent with the simulation results.

5.4 Programmable Current Source (0 ∼ 1A)

> **Relevant resources:**
>
> (1) LT1102 Data Sheet (High Speed, Precision, JFET Input Instrumentation Amplifier) (Page 9)
>
> (https://www.analog.com/media/en/technical-documentation/data-sheets/1102fb.pdf)

5.5 Current Source with High Accuracy and Fast Settling (-500mA ∼ +500mA)

> **Relevant resources:**
>
> (1) A Large Current Source with High Accuracy and Fast Settling
>
> (https://www.analog.com/media/en/analog-dialogue/volume-52/number-4/a-large-current-source-with-high-accuracy-and-fast-settling.pdf)

5.6 Other Current Source Options

(1) Current Sources: Options and Circuits (https://www.analog.com/media/en/technical-documentation/application-notes/AN-968.pdf)

Chapter 6 Voltage Source

6.1 Universal design idea

A universal design idea of a voltage-controlled voltage source (VCVS) is to increase the load capacity of the circuit from using the simplest resistor-plus-zener to adding opamp and outboard pass transistor, utill we meet the power and stability requirements. Fig.6.1 exemplifies this idea, and you can refer to *The Art of Electronics* (3rd edition, 2015) (Paul Horowitz, Winfield Hill) Page 596 for more details.

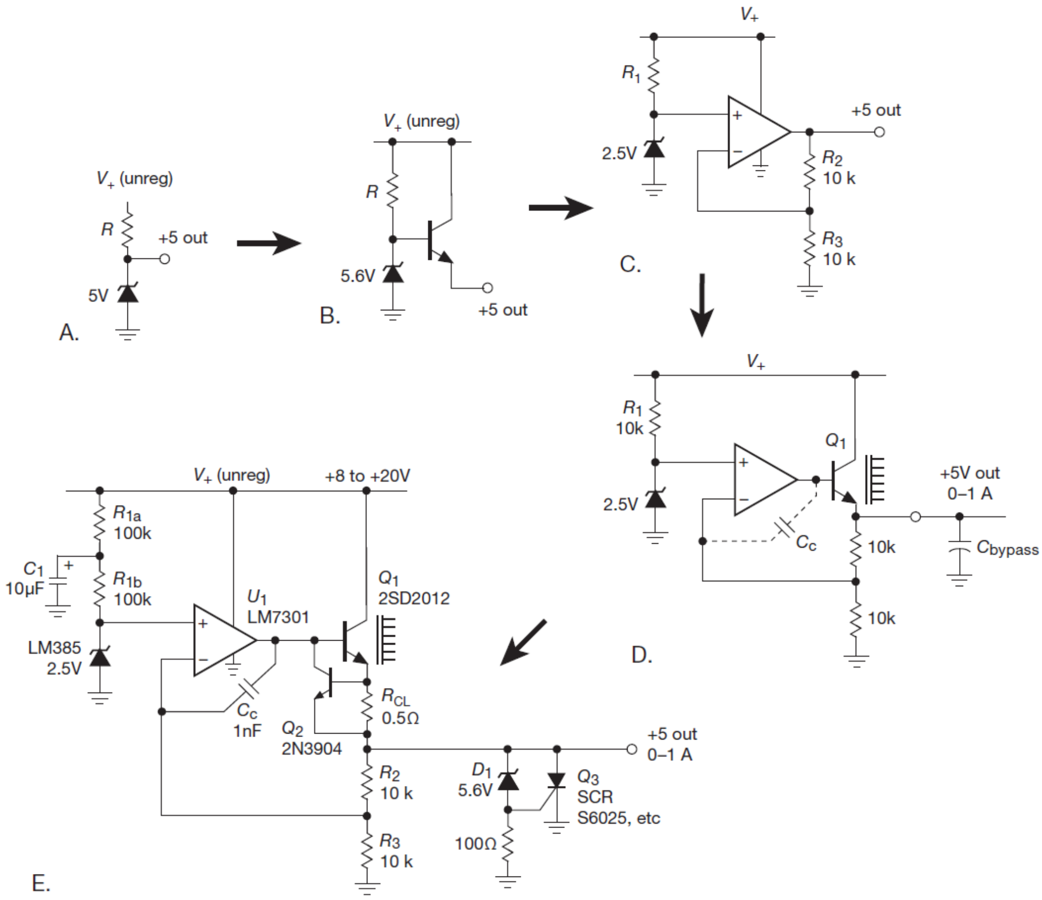


Figure 6.1: Evolving the (discrete-component) series-pass linear voltage regulator, taken from *The Art of Electronics* (3rd edition, 2015) (Paul Horowitz, Winfield Hill) Page 596

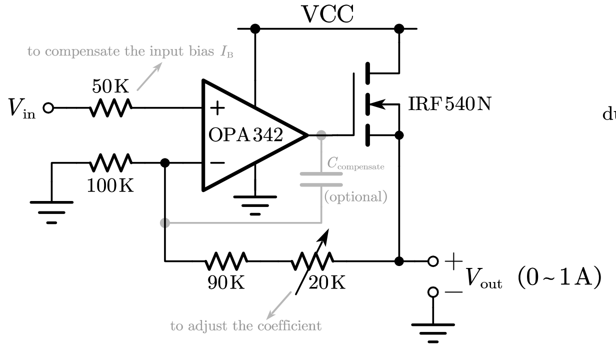
6.2 General Voltage Controlled Voltage Source (up to 5 A)

Relevant resources:

(1) *The Art of Electronics* (3rd edition, 2015) (Paul Horowitz, Winfield Hill) Page 596

6.2.1 Circuit schematic

Via the idea we mentioned in the last section, we then construct a general VCVS with 1 A load capacity (up to 5 A with additional heat sink), shown in Fig.6.2. The output voltage range is limited by VCC and GND.



Notes: single supply VCC for rail to rail opamp
dual supply $\pm V_S$ for general opamp (such as NE5532),

$V_{in} > 0$, $V_{out} = 2V_{in}$, $V_{out} < \text{VCC}$

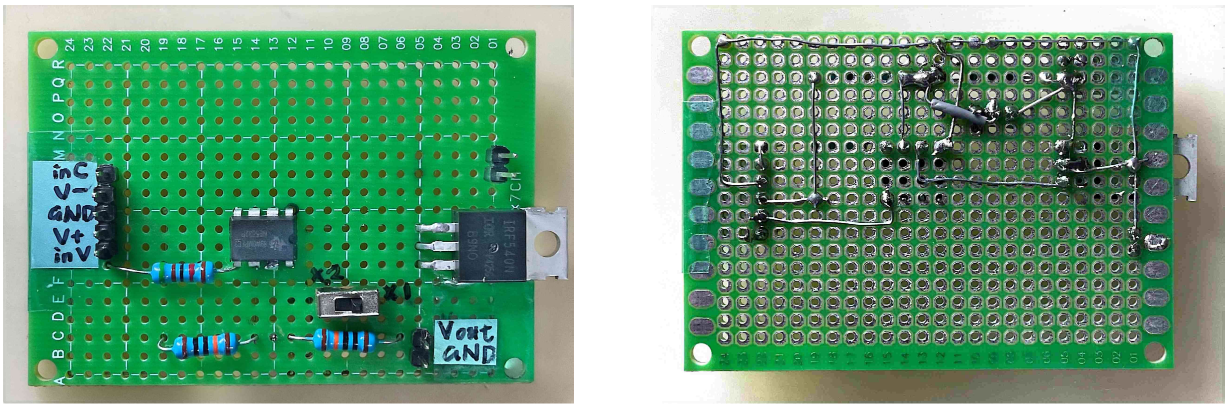
when used as a high power current source,
additional cooling devices are required

Figure 6.2: Circuit schematic for the voltage-controlled voltage source with 1 A load capacity (up to 5 A via additional heat sink)

When used as a high-power voltage source, there could be a huge power dissipation on the pass transistor (IRF540N in this case). Therefore, to avoid overheating, a heat sink is necessary for the transistor.

6.2.2 Perforated Board Verification

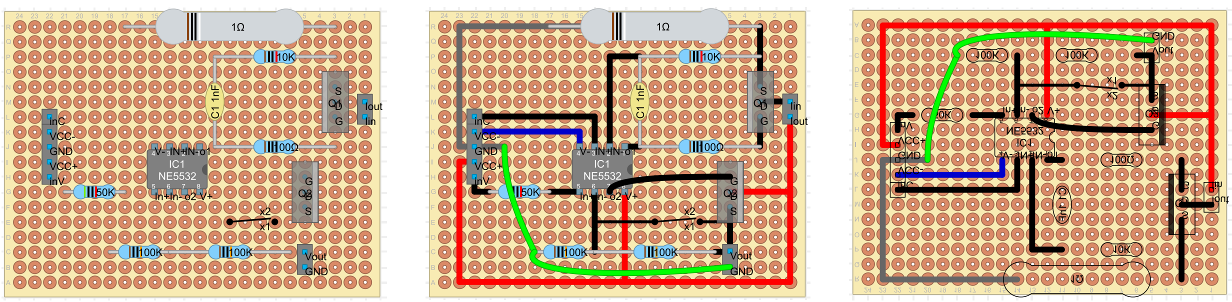
We use a general opamp NE5532P (two channels) and dual supply to construct the circuit in the perforated board as below. The switch is used to short-circuit the 100 KΩ resistor below it, obtaining an adjustable coefficient ×1 (on, 0 Ω) or ×2 (off, 100 KΩ).



(a) Top view **(b) Back view (vertically flip)**

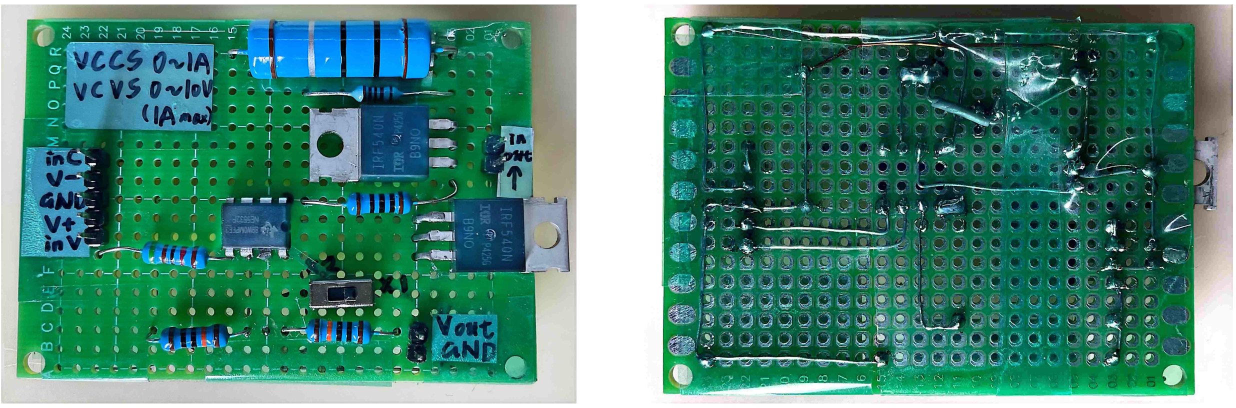
Figure 6.3: Voltage controlled voltage source (1 A max) on the perforated board

By the way, since NE5532P has two output channels, we can construct a VCCS (voltage-controlled current source) via the remaining channel to maximize the use of NE5532P. Fig.6.5 and Fig.6.5 exemplifies this idea, refer to chapter "Current Source" for a reminder.



(a) Top view **(b) Top view (with wires)** **(c) Back view (vertically flip)**

Figure 6.4: Circuit schematics for voltage-controlled current source (0 ∼ 1 A) and voltage-controlled voltage source (1 A max)



(a) Top view (with wires) **(b) Back view (vertically flip)**

Figure 6.5: Voltage-controlled current source (0 $\sim$ 1 A) and voltage-controlled voltage source (1 A max) on the perforated board

Here are the output characteristics of the circuit, measured under the condition of $V_{\mathrm{CC},\pm} = \pm 12$ V, T_A = 25 ℃ (unless otherwise noted).

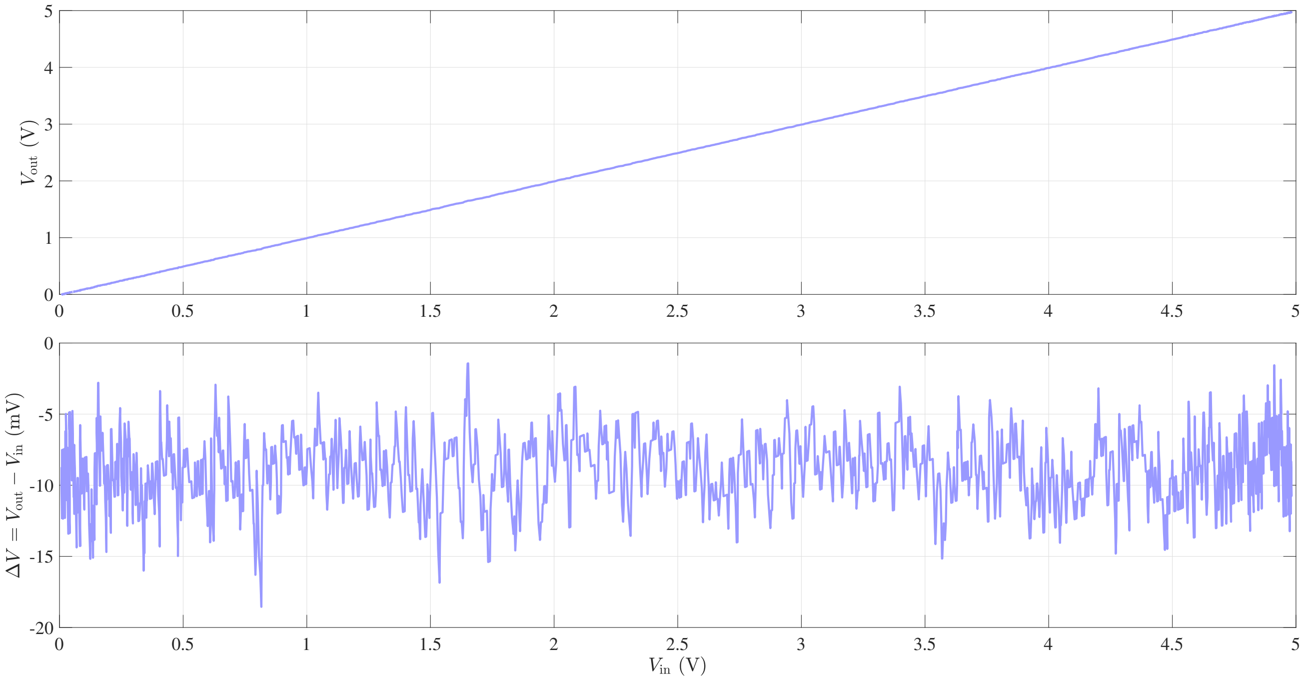


Figure 6.6: Static output characteristics, switch on (coefficient = 1, unit output), $R_L = \infty$

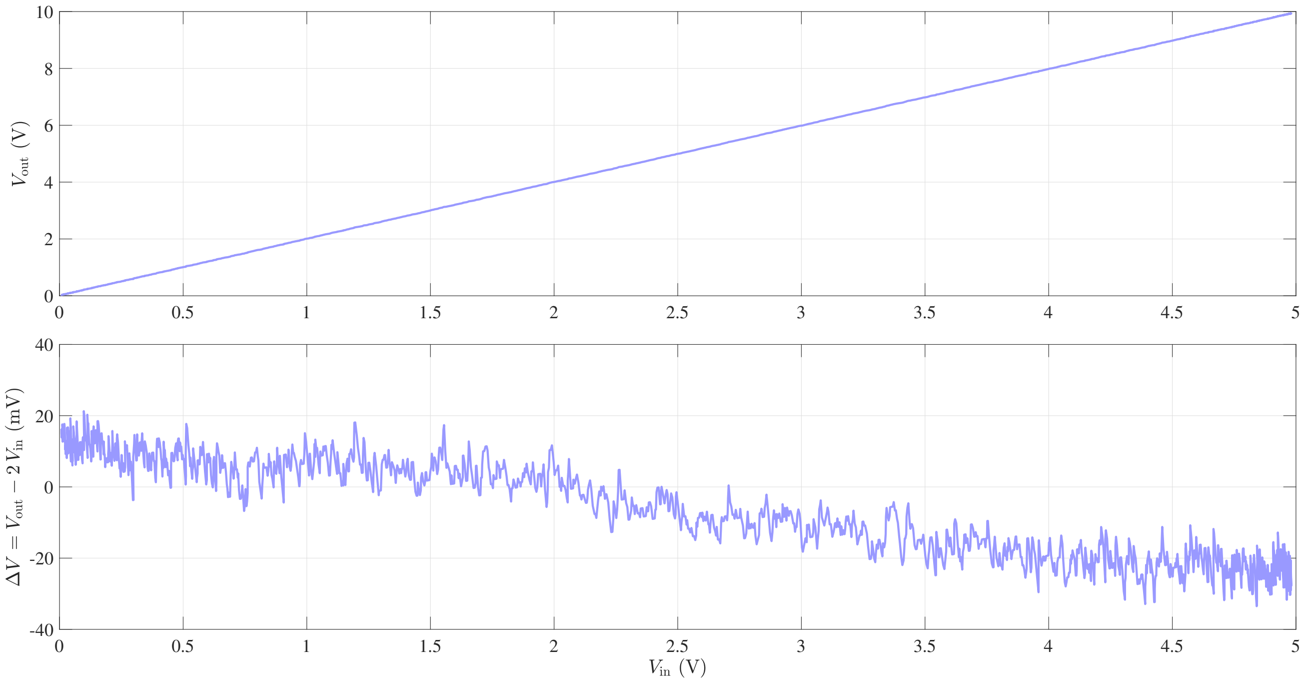


Figure 6.7: Static output characteristics, switch off (coefficient = 2, double output), $R_L = \infty$

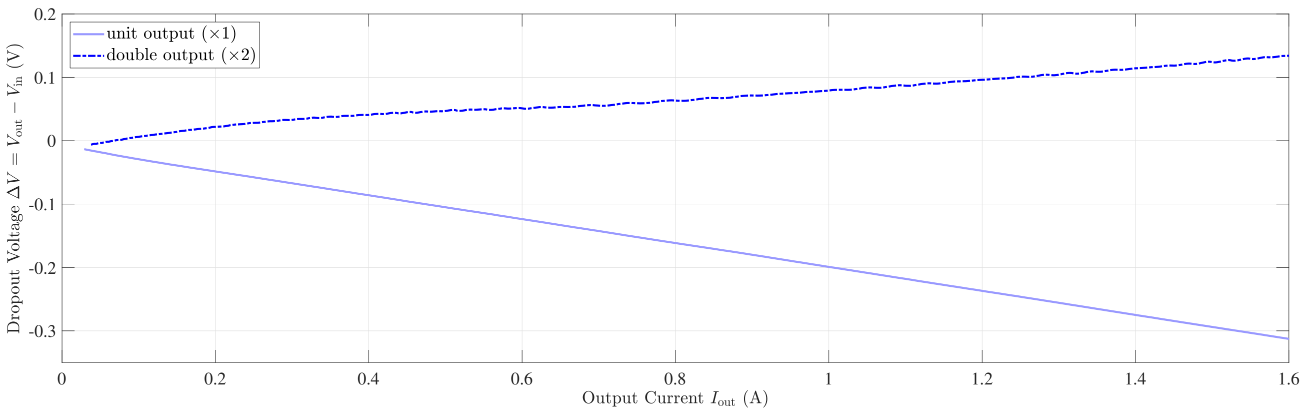


Figure 6.8: Dropout voltage ΔV vs. output current I_{out}

Note that the compensation capacitor C_c is recommanded to avoid oscillation and improve the transient response. But it is necessary for the high-frequency response usage, a suggested value range is 0.1 nF $\sim$ 1 nF. We use 0.1 nF in this case,

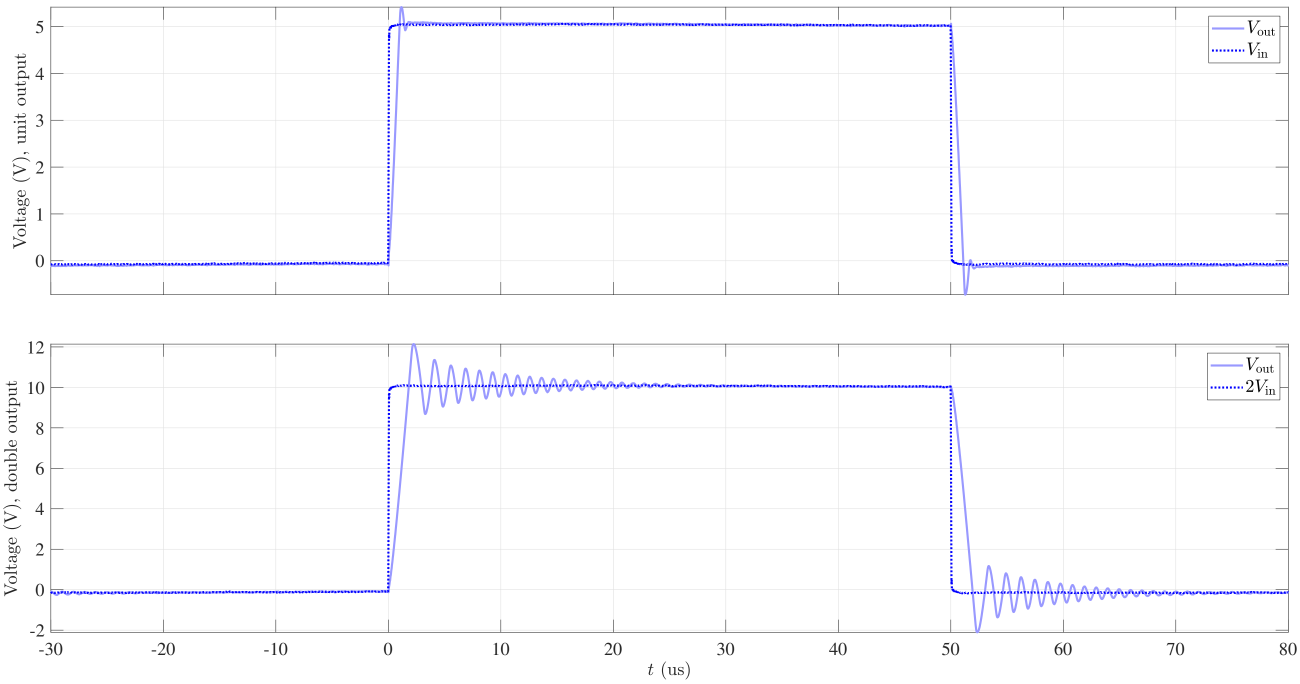


Figure 6.9: Transient response, $R_L = \infty$, $C_c = 0$

Figure 6.10: Transient response, $R_L = \infty$, $C_c = 0.1\text{nF}$

Figure 6.11: High-frequency response of unit output mode (100 KHZ input), $R_L = 100\ \Omega$, $C_c = 0.1$ nF

Figure 6.12: High-frequency response of double output mode (100 KHZ input), $R_L = 100\ \Omega$, $C_c = 0.1$ nF