RCET 2251

Systems Analog & Digital Theory

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Contents

Getting Started

0.1 Welcome

Welcome to RCET 2251: Systems Analog and Digital Theory! I'm glad to have you in this course and excited to help you explore the foundational principles that connect analog and digital systems in electronics and engineering technology. This class is designed to give you practical insight into how real-world systems operate—from analyzing analog signals and digital logic to understanding system behavior, feedback, and control.

Whether you're working with amplifiers, filters, timing circuits, or embedded systems, the concepts we cover will give you the tools to model, analyze, and troubleshoot complex systems with confidence. We'll focus on making these topics approachable, hands-on, and directly relevant to your future work in the field.

Bring your curiosity, ask questions, and get ready to link theory with real-world applications. Let's dive in!

0.2 Recommended Books

Recommended Books:

- RCET First and Second Semester Text Books and Lab Reports.
- *Solid State Pulse Circuits* David A. Bell [2]

0.3 Syllabus & RCET Student Handbook

- RCET 2251 Syllabus
- RCET Student Handbook

Week 1

Oscilloscope & Test Equipment Familiarization

1.1 Objectives

Review Bench Equipment Specifications, General Operation, and Test and Measurement Procedures as Identified in the Operator's Manual.

1. **Equipment Specifications:** Students should be able to review and understand the specifications of bench equipment, including details such as voltage ranges, current limits, frequency response, and other relevant parameters as outlined in the operator's manual.

2. **General Operation:** Gain knowledge of the general operation principles of bench equipment, understanding functions, controls, and interface features described in the operator's manual.

3. **Test and Measurement Procedures:** Familiarize themselves with the recommended test and measurement procedures provided in the operator's manual for various bench equipment. This includes understanding how to perform accurate measurements, set appropriate parameters, and interpret results.

4. **Understand and Identify Different Types and Characteristics of Waveforms.**

    - Students should be able to identify and understand different types of waveforms, including periodic and aperiodic waveforms.

5. **Understand Frequency Synthesis and the Process of Combining Sine Waves.**

    - Gain insight into the process of combining sine waves to produce complex waveforms. Understand the principles of frequency synthesis and its application.

By achieving this theory objective, students will develop a solid theoretical foundation regarding the specifications, operation, and testing procedures for bench equipment. This knowledge is crucial for the effective and accurate utilization of the equipment in practical applications.

1.2 Reference Documents:

Tektronix XYZs of Oscilloscopes PDF [3]

Tektronix TDS Oscilloscope User Manual PDF [4]

*Tektronix AFG1022 Arbitrary/Function Generator Quick Start User Manual*PDF [5]

*Power Supply GPS-4303 User Manual*PDF [6]

Tektronix - Get more from your basic oscilloscope with the FFT function [7]

1.3 Tektronix XYZs of Oscilloscopes Part I

Review the Tektronix XYZs of Oscilloscopes and answer the following section questions:

1.3.1 Match the following:

1. ____ Acquisition
2. ____ Analog
3. ____ Bandwidth
4. ____ Digital Phosphor
5. ____ Frequency
6. ____ Glitch
7. ____ Period
8. ____ Phase
9. ____ Pulse
10. ____ Waveform Point
11. ____ Rise Time
12. ____ Sample Point
13. ____ Digital Storage
14. ____ Time Base
15. ____ Transient
16. ____ ADC Resolution
17. ____ Volt

A The unit of electric potential difference.

B A performance measurement indicating the precision of an ADC, measured in bits.

C Term used when referring to degree points of a signal's period.

D The number of times a signal repeats in one second.

E The amount of time it takes a wave to complete one cycle.

F A stored digital value that represents the voltage of a signal at a specific point in time on the display.

G A common waveform shape that has a rising edge, a width, and a falling edge.

H A performance measurement indicating the rising edge speed of a pulse.

I Oscilloscope circuitry that controls the timing of the sweep.

J An intermittent spike in a circuit.

K A signal measured by an oscilloscope that only occurs once.

L The oscilloscope's process of collecting sample points from the ADC, processing them, and storing them in memory.

M Something that operates with continuously changing values.

N Digital oscilloscope that captures 3 dimensions of signal information in real-time.

O Digital oscilloscope with serial processing.

P A sine wave frequency range, defined by the -3dB point.

Q The raw data from an ADC used to calculate and display waveform points.

1.3.2 Multiple Choice:

1. With an oscilloscope you can:
   (a) Calculate the frequency of a signal.
   (b) Find malfunctioning electrical components.
   (c) Analyze signal details.
   (d) All the above.

2. The difference between analog and digitizing oscilloscopes is:
   (a) Analog oscilloscopes do not have on-screen menus.
   (b) Analog oscilloscopes apply a measurement voltage directly to the display system, while digital oscilloscopes first convert the voltage into digital values.
   (c) Analog oscilloscopes measure analogs, whereas digitized oscilloscopes measure digits.
   (d) Analog oscilloscopes do not have an acquisition system.

3. An oscilloscope's vertical section does the following:
   (a) Acquires sample points with an ADC.
   (b) Starts a horizontal sweep.
   (c) Lets you adjust the brightness of the display.
   (d) Attenuates or amplifies the input signal.

4. The time base control of the oscilloscope does the following:
   (a) Adjusts the vertical scale.
   (b) Shows you the current time of day.
   (c) Sets the amount of time represented by the horizontal width of the screen.
   (d) Sends a clock pulse to the probe.
5. On an oscilloscope display:
   (a) Voltage is on the vertical axis and time is on the horizontal axis.
   (b) A straight diagonal trace means voltage is changing at a steady rate.
   (c) A flat horizontal trace means voltage is constant.
   (d) All the above.
6. All repeating waves have the following properties:
   (a) A frequency measured in Hertz.
   (b) A period measured in seconds.
   (c) a bandwidth measured in Hertz.
   (d) All the above.
7. If you probe inside a computer with an oscilloscope, you are likely to find the following types of signals:
   (a) Pulse trains.
   (b) Ramp waves.
   (c) Sine waves.
   (d) All the above.
8. When evaluating the performance of an analog oscilloscope, some things you might consider are:
   (a) The bandwidth.
   (b) The vertical sensitivity.
   (c) The ADC resolution.
   (d) The sweep speed.
9. The difference between digital storage oscilloscopes (DSO) and digital phosphor oscilloscopes (DPO) is:
   (a) The DSO has a higher bandwidth.
   (b) The DPO captures three dimensions of waveform information in real-time.
   (c) The DSO has a color display.
   (d) The DSO captures more signal details.

1.4 Tektronix XYZs of Oscilloscopes Part II

Review the Tektronix XYZs of Oscilloscopes and answer the following section questions:

1.4.1 Match the following:

1. ____ Averaging Mode
2. ____ Circuit Loading
3. ____ Compensation
4. ____ Coupling
5. ____ Earth Ground
6. ____ Equivalent-Time
7. ____ Graticule
8. ____ Interpolation
9. ____ Real Time
10. ____ Signal Generator
11. ____ Single Sweep
12. ____ Sensor

A The unintentional interaction of the probe and oscilloscope with the circuit being tested which distorts a signal.

B A conductor that connects electrical currents to the Earth.

C A sampling mode in which the digital oscilloscope collects as many samples as it can as the signal occurs, then constructs a display, using interpolation if necessary.

D A sampling mode in which the digital oscilloscope constructs a picture of a repetitive signal by capturing a little bit of information from each repetition.

E A device that converts a specific physical quantity such as sound, pressure, strain, or light intensity into an electrical signal.

F A test device for injecting a signal into a circuit input.

G A processing technique used by digital oscilloscopes to eliminate noise in a displayed signal.

H The method of connecting two circuits together.

I A "connect-the-dots" processing technique to estimate what a fast waveform looks like based on only a few sampled points.

J The grid lines on a screen for measuring oscilloscope traces.

K A trigger mode that triggers the sweep once, must be reset to accept another trigger event.

L A probe adjustment for 10X attenuator probes that balances the electrical properties of the probe with the electrical properties of the oscilloscope.

1.4.2 Multiple Choice:

1. To operate an oscilloscope safely, you should:
   (a) Ground the oscilloscope with the proper three-pronged power cord.
   (b) Learn to recognize potentially dangerous electrical components.
   (c) Avoid touching exposed connections in a circuit being tested even if the power is off.
   (d) All the above.

2. Grounding an oscilloscope is necessary:
   (a) For safety reasons.
   (b) To provide a reference point for making measurements.
   (c) To align the trace with the screen's horizontal axis.
   (d) All the above.

3. Circuit loading is caused by:
   (a) An input signal having too large a voltage.
   (b) The probe and oscilloscope interacting with the circuit being tested.
   (c) a 10X attenuator probe being uncompensated.
   (d) Putting too much weight on a circuit.

4. Compensating a probe is necessary to:
   (a) Ballance the electrical properties of the 10X attenuator probe with the oscilloscope.
   (b) Prevent damaging the circuit being tested.
   (c) Improve the accuracy of your measurements.
   (d) All the above.

5. The trace rotation control is useful for:
   (a) Scaling waveforms on the screen.
   (b) Detecting sine wave signals.
   (c) Aligning the waveform trace with the screen's horizontal axis on an analog oscilloscope.
   (d) Measuring pulse width.

6. The volts per division control is used to:(select all that apply)
   - ☐ Scale a waveform vertically.
   - ☐ Position a waveform vertically.
   - ☐ Attenuate or amplify an input signal.
   - ☐ Set the number of volts each division represents.
7. Setting the vertical input coupling to ground does the following:
   - ☐ Disconnects the input signal from the oscilloscope.
   - ☐ Causes a horizontal line to appear with auto trigger.
   - ☐ Lets you see where zero volts is on the screen.
   - ☐ All the above.
8. The trigger is necessary to:
   - ☐ Stabilize repeating waveforms on the screen.
   - ☐ Capture single-shot waveforms.
   - ☐ Mark a particular point of an acquisition.
   - ☐ All the above.
9. The difference between auto and normal trigger mode is:
   - ☐ In normal mode the oscilloscope only sweeps once and then stops.
   - ☐ In normal mode the oscilloscope only sweeps if the input signal reaches the trigger point; otherwise the screen is blank.
   - ☐ Auto mode makes the oscilloscope sweep continuously even without being triggered.
   - ☐ All the above.
10. The acquisition mode that best reduces noise in a repeating signal is:
    - ☐ Sample mode.
    - ☐ Peak detect mode.
    - ☐ Envelope mode.
    - ☐ Averaging mode.

11. The two most basic measurements you can make with an oscilloscope are:

    - ☐ Time and frequency measurements.
    - ☐ Time and voltage measurements.
    - ☐ Voltage and pulse width measurements.
    - ☐ Pulse width and phase shift measurements.

12. If the volts/division is set at 0.5, the largest signal that can fit on the screen (assuming an 8 x 10 division screen) is:

    - ☐ 62.5 millivolts peak-to-peak.
    - ☐ 8 volts peak-to-peak.
    - ☐ 4 volts peak-to-peak.
    - ☐ 0.5 volts peak-to-peak.

13. If the seconds/division is set at 0.1 ms, the amount of time represented by the width of the screen is:

    - ☐ 0.1 ms.
    - ☐ 1 ms.
    - ☐ 1 second.
    - ☐ 0.1 kHz.

14. By convention, pulse width is measured:

    - ☐ At 10% of the pulse's peak-to-peak (pk-pk) voltage.
    - ☐ At 50% of the pulse's peak-to-peak (pk-pk) voltage.
    - ☐ At 90% of the pulse's peak-to-peak (pk-pk) voltage.
    - ☐ At 10% and 90% of the pulse's peak-to-peak (pk-pk) voltage.

15. You attach a probe to your test circuit but the screen is blank. You should:

    - ☐ Check that the screen intensity is turned up.
    - ☐ Check that the oscilloscope is set to display the channel that the probe is connected to.
    - ☐ Set the trigger mode to auto since norm mode blanks the screen.
    - ☐ Set the vertical input coupling to AC and set the volts/division to its largest value since a large DC signal may go off the top or bottom of the screen.
    - ☐ Check that the probe isn't shorted and make sure it is properly grounded.
    - ☐ Check that the oscilloscope is set to trigger on the input channel you are using.
    - ☐ All of the above.

1.5 Probe Compensation

Review the Tektronix XYZs of Oscilloscopes and answer the following section questions:

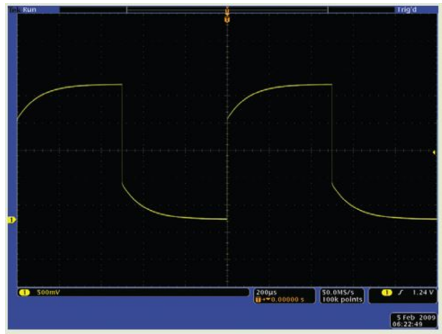


Figure 1.1: Probe Adjustment Signal

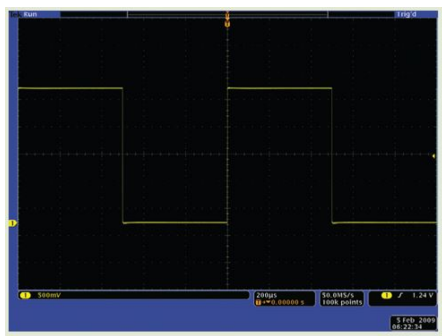


Figure 1.2: Probe Adjustment Signal

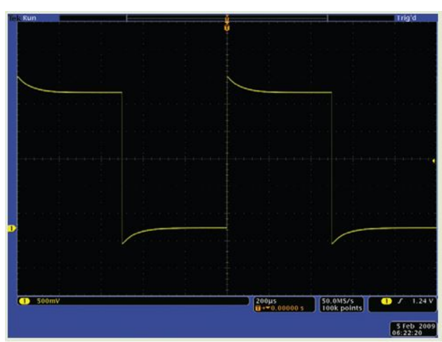


Figure 1.3: Probe Adjustment Signal

1. The Probe Adjustment Signal as seen in **Figure 1.1** represents a:

   - ☐ under-compensated probe.
   - ☐ properly compensated probe.
   - ☐ over-compensated probe.

2. The Probe Adjustment Signal as seen in **Figure 1.2** represents a:

   - ☐ under-compensated probe.
   - ☐ properly compensated probe.
   - ☐ over-compensated probe.

3. The Probe Adjustment Signal as seen in **Figure 1.3** represents a:

   - ☐ under-compensated probe.
   - ☐ properly compensated probe.
   - ☐ over-compensated probe.

1.6 AFG1022

Review the Tektronix AFG1022 Quick Start User Manual and answer the following section questions:

1. The AFG1022 can produce which of the following waveforms:
   - ☐ Sine
   - ☐ Square
   - ☐ Ramp
   - ☐ Pulse
   - ☐ Noise
2. The Output impedance of the AFG1022 is ____________ .
   - ☐ 50Ω
   - ☐ 600Ω
   - ☐ $1K\Omega$
   - ☐ $1M\Omega$
   - ☐ $10M\Omega$
3. Push the front-panel ____________ button to control the screen display. You can toggle between the two channels.
   - ☐ On/Off
   - ☐ Both
   - ☐ Ch1
   - ☐ Ch2
   - ☐ Ch1/2
4. To enable CH1 signal output, push the front-panel On/Off with ____________ color.
   - ☐ Red
   - ☐ Green
   - ☐ Blue
   - ☐ Yellow

5. To enable CH2 signal output, push the front-panel On/Off with ____________ color.

   - ☐ Red
   - ☐ Green
   - ☐ Blue
   - ☐ Yellow

6. The Sweep outputs a waveform with the output signal frequency varying ____________ or ____________ .

   - ☐ continuously, non-continuously
   - ☐ linearly, logarithmically
   - ☐ synchronized, non-synchronized
   - ☐ sweep, digital pulse sweep

7. According to the AFG1022 Quick Start User Manual, What fuse should be used?

   - ☐ 250,F0.5AL
   - ☐ 250,F1AL
   - ☐ 250,F1.5AL
   - ☐ 250,F2AL

1.7 Graphs and Waveforms

1.7.1 What is a Graph?

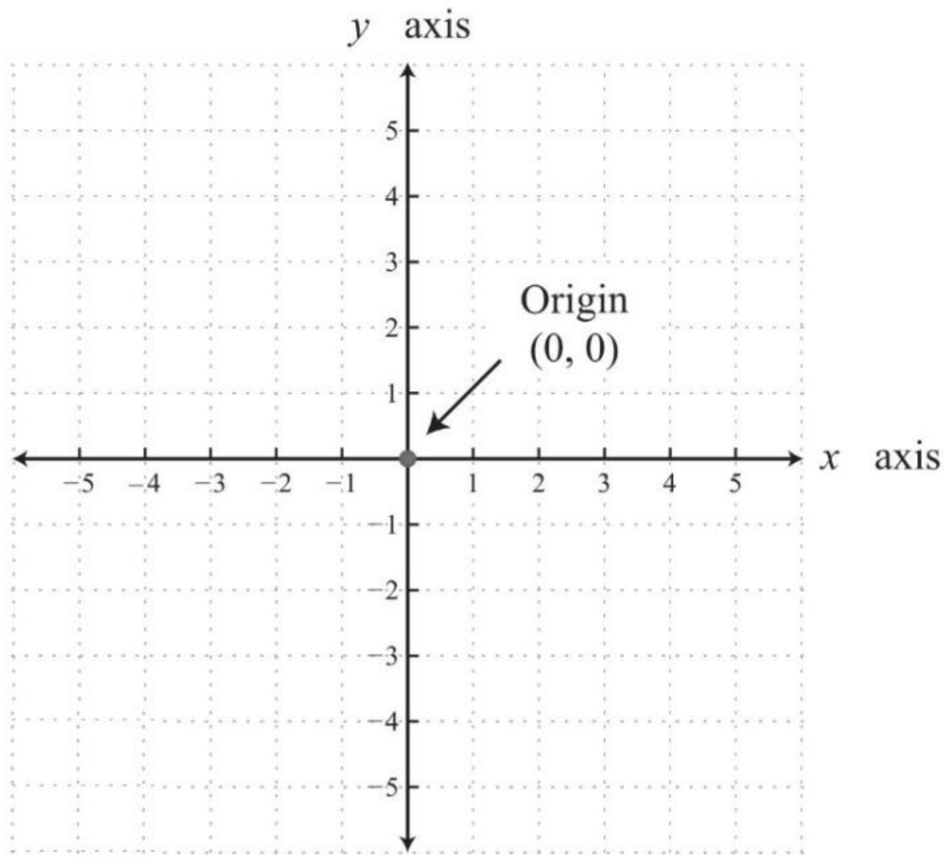


Figure 1.4: Graph

- **Horizontal Axis (X-Axis):** This axis runs horizontally from left to right and is usually used to represent the independent variable or input.
- **Vertical Axis (Y-Axis):** This axis runs vertically from bottom to top and is usually used to represent the dependent variable or output.
- The point where the two axes intersect is called the **Origin**. The coordinates of the origin are typically represented as (0,0). The horizontal and vertical distances from the origin to any point on the graph are called the x-coordinate and y-coordinate, respectively. The combination of these coordinates uniquely identifies a point in the coordinate system. This system is fundamental in graphing and visualizing mathematical functions and relationships.

1.7.2 Periodic Waveforms

A **periodic waveform** is a type of waveform that repeats its shape over regular intervals of time. In other words, the waveform exhibits a regular and predictable pattern, and the pattern repeats itself after a specific period. The time it takes for one complete cycle to occur is called the period.

Key characteristics of periodic waveforms include:

1. **Frequency (f):** The frequency of a periodic waveform is the number of cycles that occur in one second. It is the reciprocal of the period and is measured in Hertz (Hz).

$$Frequency = \frac{1}{Time}$$

2. **Amplitude:** The amplitude of a periodic waveform represents the maximum displacement from the equilibrium position. It is a measure of the strength or intensity of the waveform. Amplitude is typically measured in *volts peak* or *volts peak to peak* using an oscilloscope.

3. **Phase:** The phase of a periodic waveform indicates the relative position of the waveform at a specific point in time within its cycle. It is often measured in degrees or radians.

Common examples of periodic waveforms include:

- **Sine Wave:** A smooth, oscillating waveform characterized by its sinusoidal shape.
- **Square Wave:** A waveform that alternates between two constant levels, resembling a square shape.
- **Triangular Wave:** A waveform that linearly increases and decreases in amplitude, forming a triangular shape.
- **Ramp Wave:** A waveform characterized by a linear increase or decrease in amplitude over time.
- **Sawtooth Wave:** A waveform that rises linearly and then falls abruptly, creating a sawtooth-like pattern.
- **Exponential Wave:** A waveform in which the amplitude changes exponentially over time.
- **Spike Wave:** A waveform characterized by a sudden, brief increase in amplitude, often appearing as a sharp spike in the signal.

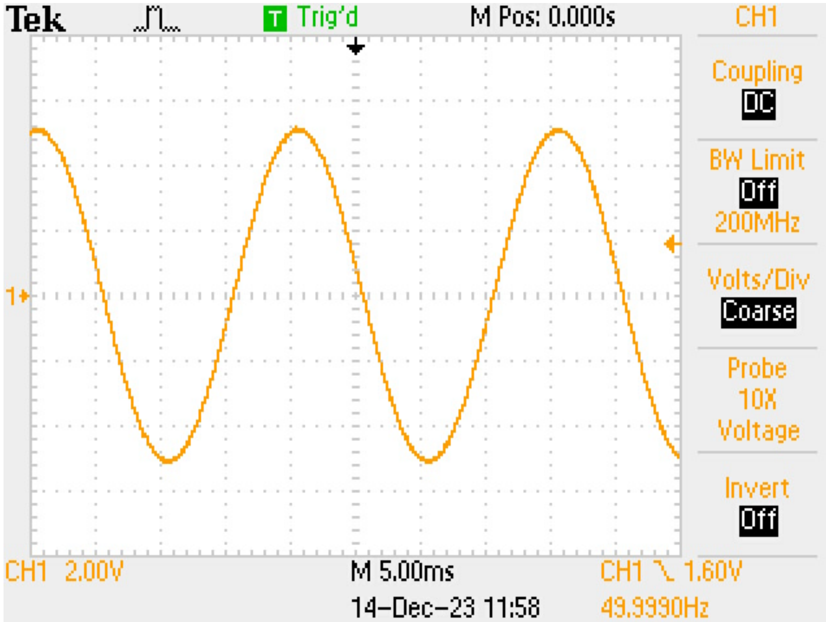


Figure 1.5: Sine Wave

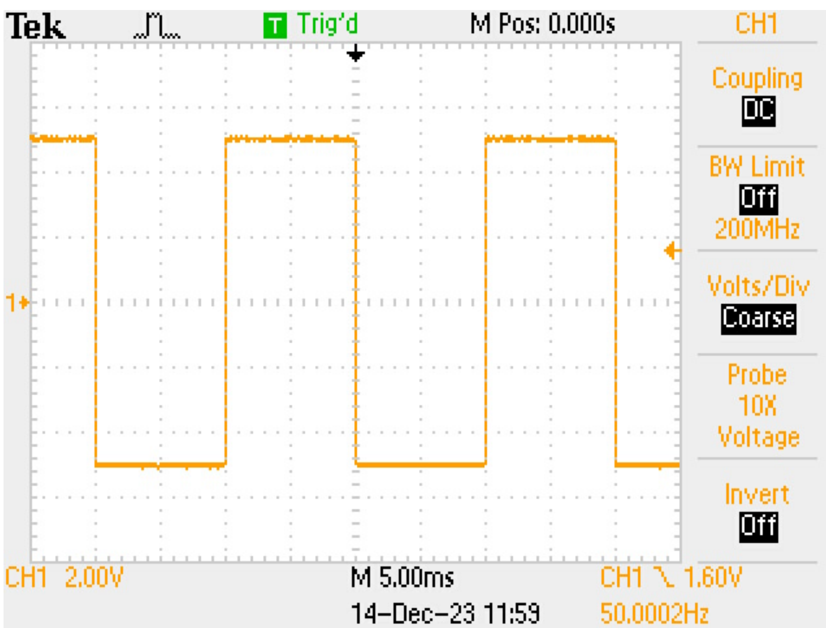


Figure 1.6: Square Wave

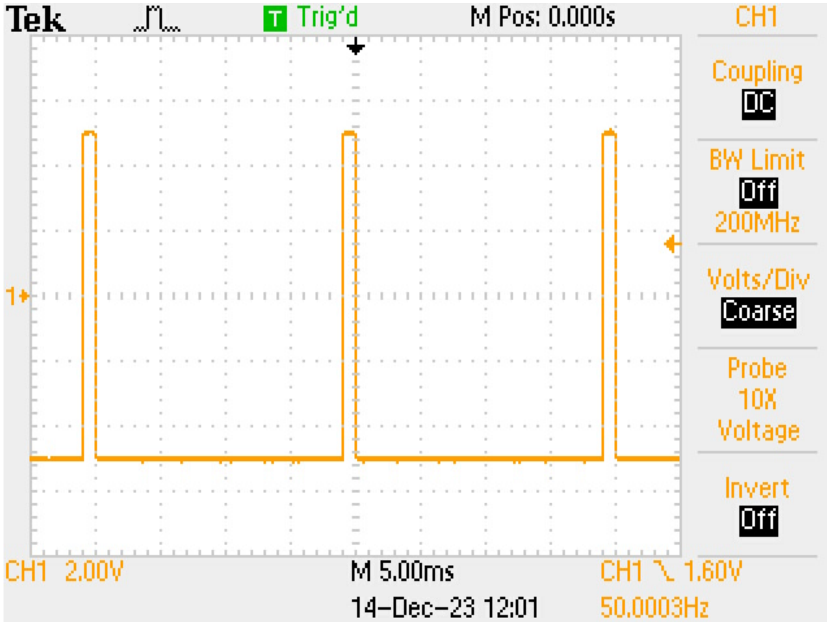


Figure 1.7: Pulse Wave

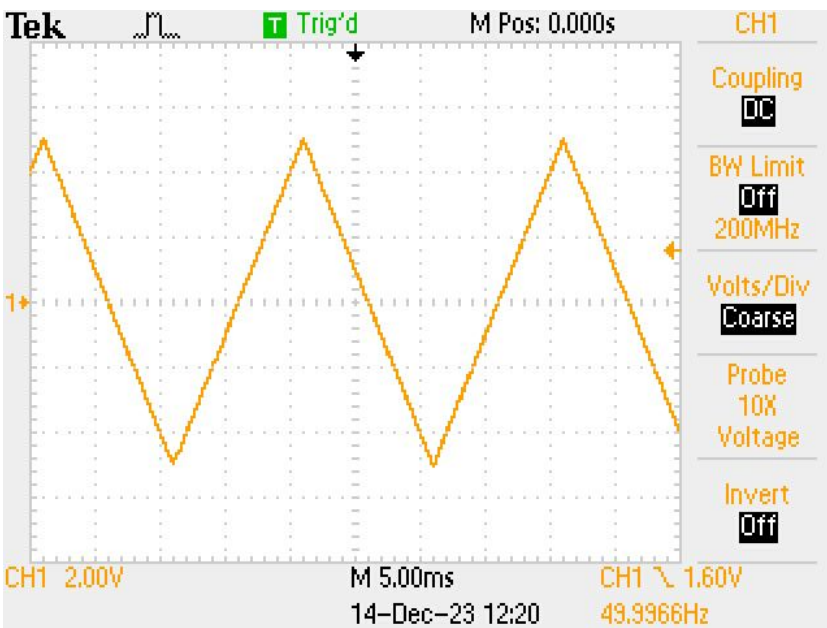


Figure 1.8: Triangle Wave

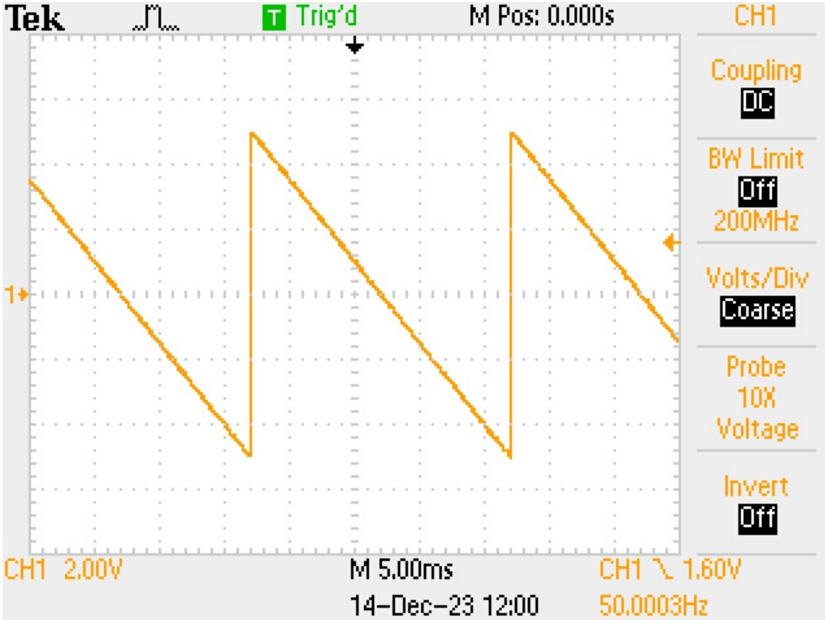


Figure 1.9: Sawtooth Wave

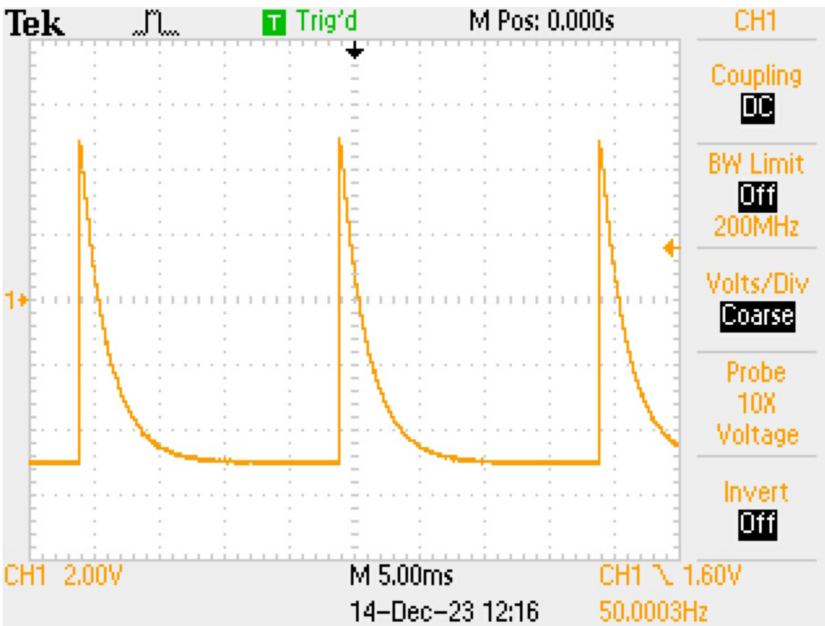


Figure 1.10: Exponential Spike Wave

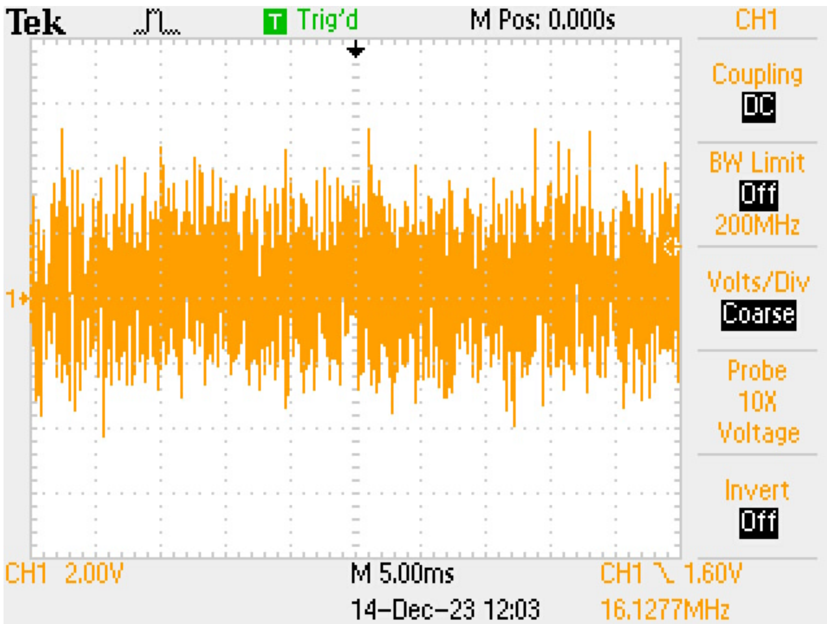


Figure 1.11: Noise Wave

1.7.3 Aperiodic Waveforms

Aperiodic waveforms, in contrast to periodic waveforms, do not exhibit a regular and repeating pattern over time. These waveforms lack a well-defined period, and their shapes do not recur in a predictable manner. Aperiodic waveforms are often associated with transient or non-repetitive signals.

Examples and characteristics:

1. **Impulse or Spike:** A waveform characterized by an instantaneous, brief increase in amplitude, representing an abrupt change.
2. **Step Function:** A waveform that changes abruptly from one constant level to another, creating a step-like pattern.
3. **Random Noise:** A waveform characterized by random variations in amplitude over time, without a discernible pattern.
4. **Exponential Decay:** A waveform where the amplitude decreases exponentially over time.
5. **Chirp Signal:** A waveform with a continuously changing frequency, often used in radar and communication systems.
6. **Pulse Train:** A series of individual pulses, where the intervals between pulses may not be constant.

1.8 Frequency Synthesis & Analysis

1.8.1 Frequency Synthesis

Frequency synthesis involves combining two or more signals to create a new waveform. Just like an AC signal will superimpose on a DC signal, two AC signals present at the same time will also combine to form a new waveform. When this occurs it is referred to as Frequency Synthesis.

> **synthesis** noun: the composition or combination of parts or elements so as to form a whole. Merriam Webster

> **Frequency Synthesis** is the process of combining multiple sine waves to produce a new desired waveform.

Harmonics

- ✓ A harmonic is a multiple of the fundamental.
- ✓ Harmonics are numbered according to their ratio to the fundamental.
- ✓ The number of harmonics is infinite; however, the amplitude of each harmonic will successively decrease as frequency increases.

Table 1.1: Sinusoidal Harmonics

Harmonic Number	Frequency	Amplitude
Fundamental	F	V
2^{nd} harmonic	2F	$\frac{V}{2}$
3^{rd} harmonic	3F	$\frac{V}{3}$
4^{th} harmonic	4F	$\frac{V}{4}$
5^{th} harmonic	5F	$\frac{V}{5}$
6^{th} harmonic	6F	$\frac{V}{6}$
7^{th} harmonic	7F	$\frac{V}{7}$
100^{th} harmonic	100F	$\frac{V}{100}$

1.8.2 Perfect Square Waves

A **Perfect Square Wave** is comprised of an infinite number of odd harmonic sine waves.

Sawtooth, Exponential, and Triangle Waveforms are comprised of a combination of odd and even harmonic sine waves.

Can a Square Wave really be created using Sine Wave Frequency Synthesis?

To test this theory we can use Desmos Graphing Calculator [8]. Link to final graph https://www.desmos.com/calculator/bbjnjmrxyz.

Sine-wave formulas:

$$Fundamental = \frac{4}{\pi}\frac{1}{1}\sin(1\pi x)$$

$$3^{rd} harmonic = \frac{4}{\pi}\frac{1}{3}\sin(3\pi x)$$

$$5^{th} harmonic = \frac{4}{\pi}\frac{1}{5}\sin(5\pi x)$$

$$7^{th} harmonic = \frac{4}{\pi}\frac{1}{7}\sin(7\pi x)$$

$$9^{th} harmonic = \frac{4}{\pi}\frac{1}{9}\sin(9\pi x)$$

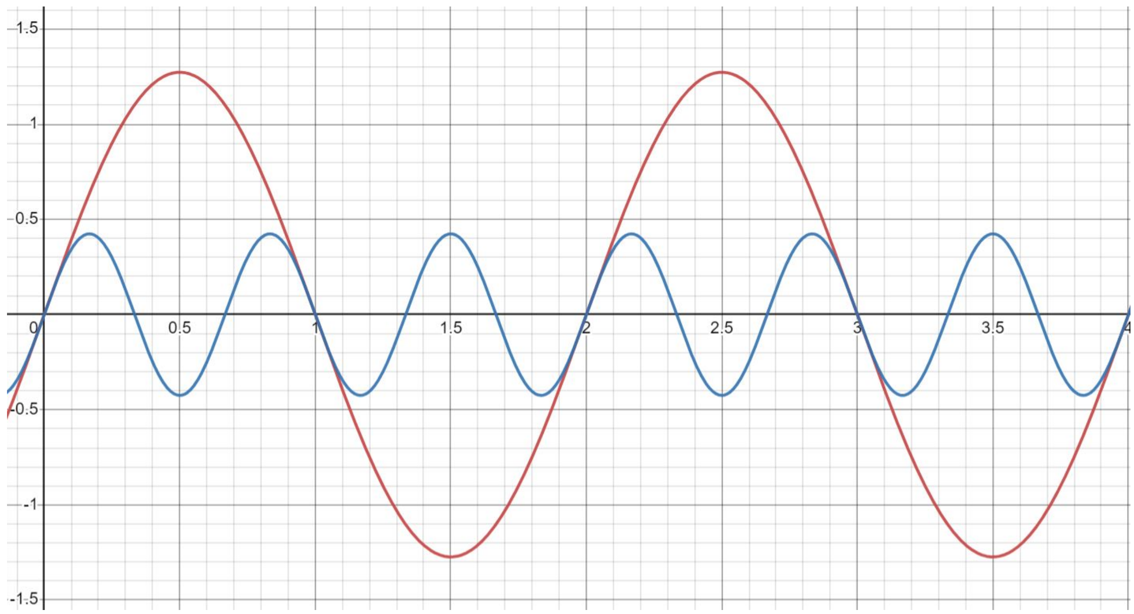


Figure 1.12: Fundamental & Third Harmonic

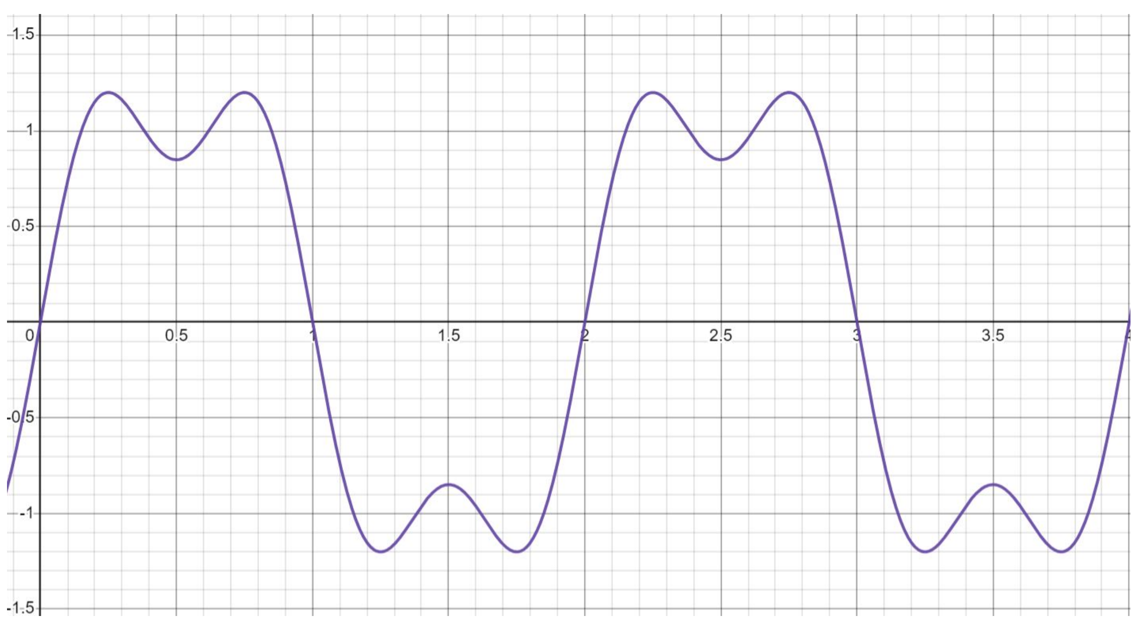


Figure 1.13: Fundamental & Third Harmonic Synthesis

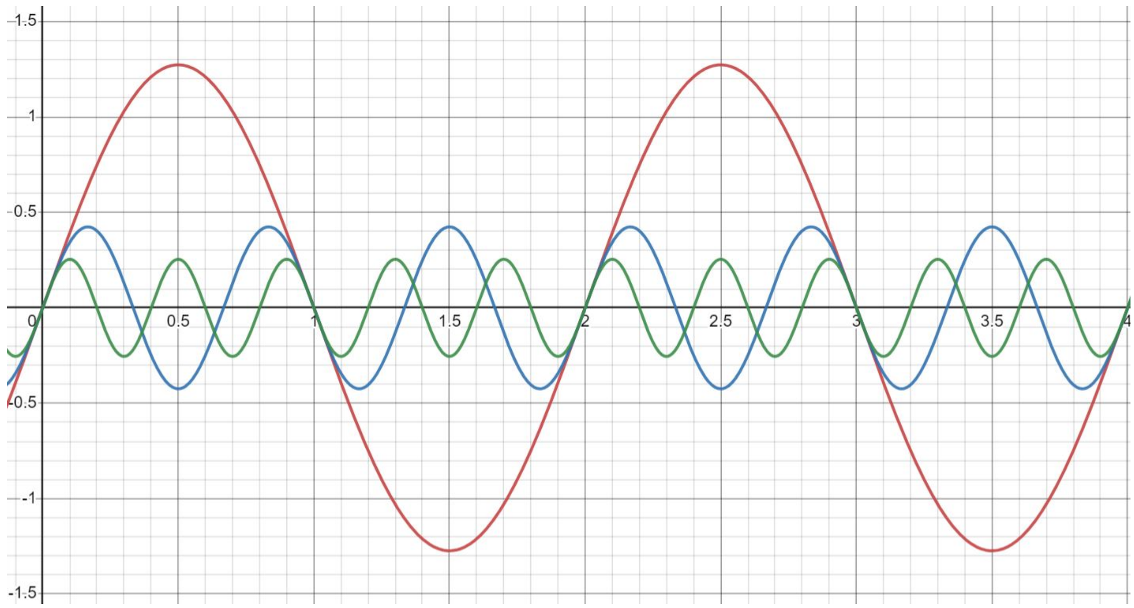


Figure 1.14: Fundamental, Third & Fifth Harmonics

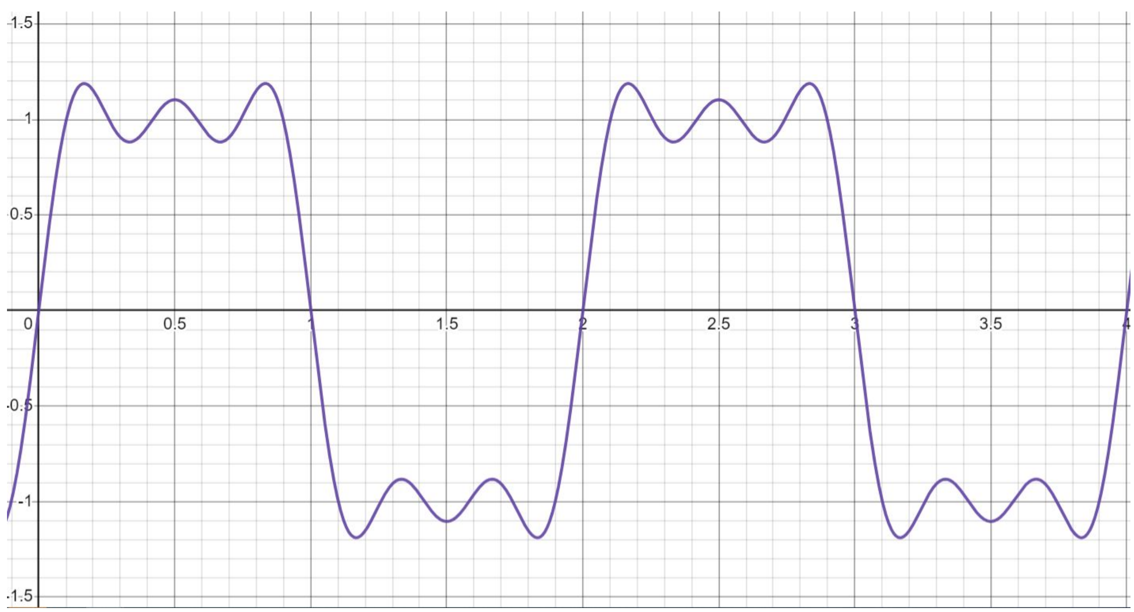


Figure 1.15: Fundamental, Third & Fifth Harmonics Synthesis

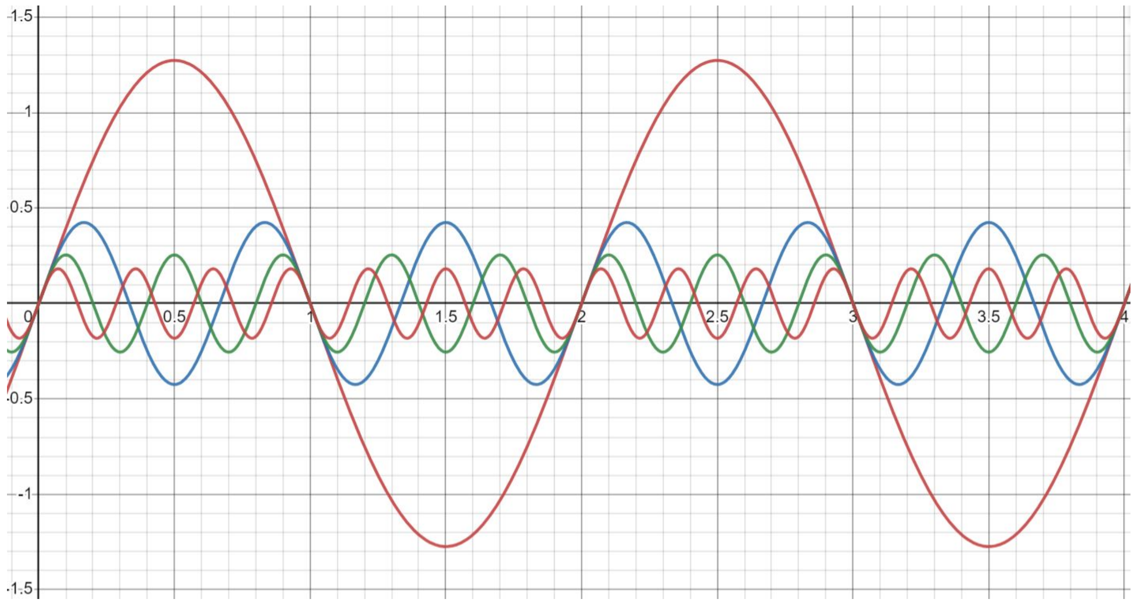


Figure 1.16: Fundamental, Third, Fifth, & Seventh Harmonics

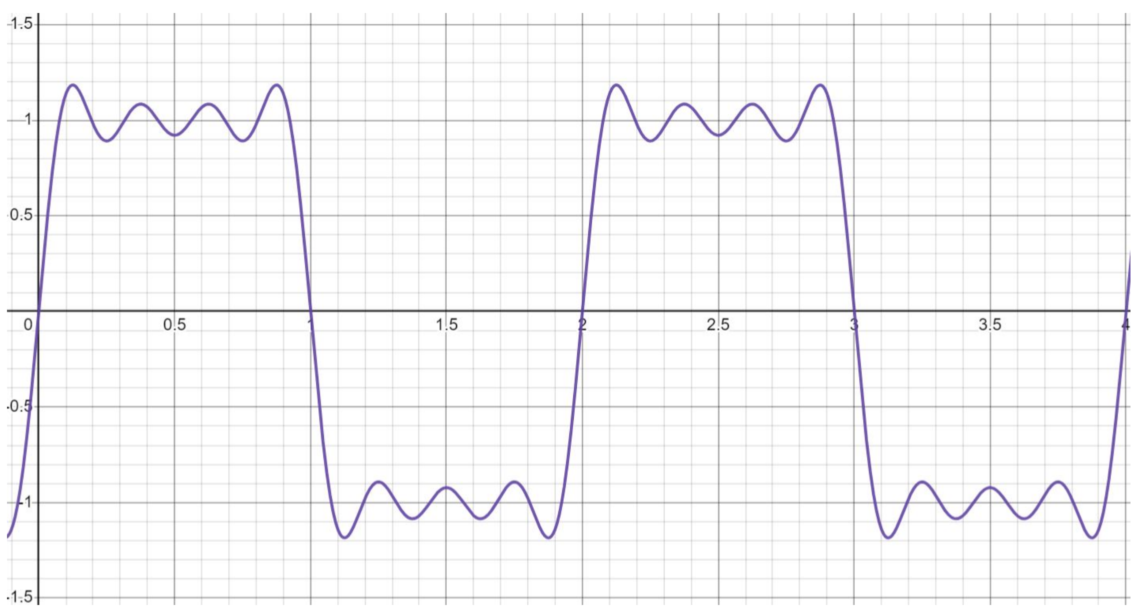


Figure 1.17: Fundamental, Third, Fifth, & Seventh Harmonics Synthesis

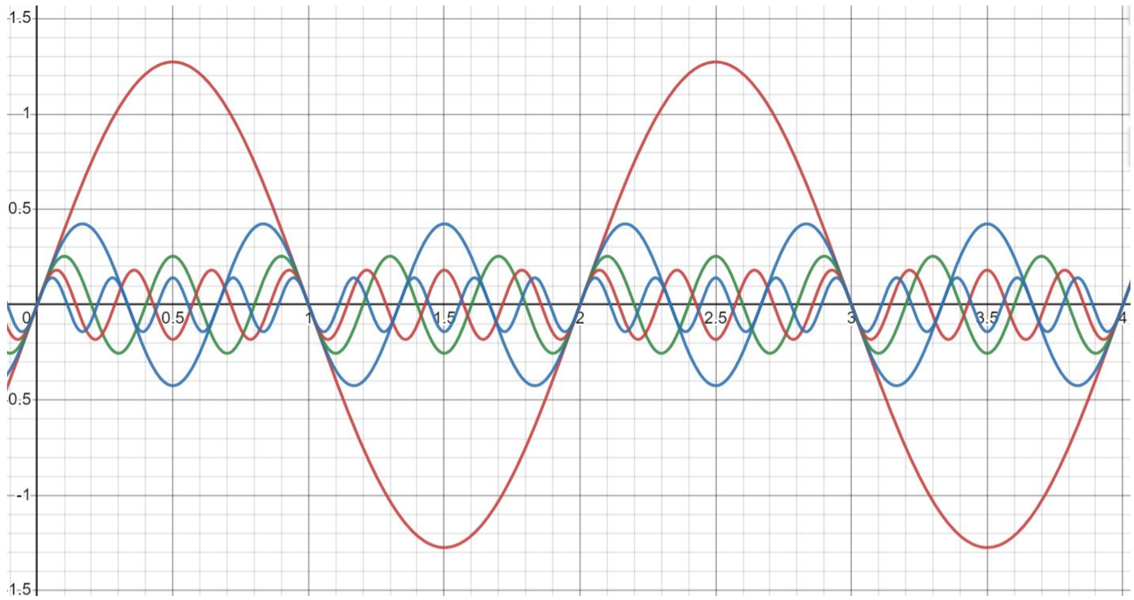


Figure 1.18: Fundamental, Third, Fifth, Seventh, & Ninth Harmonics

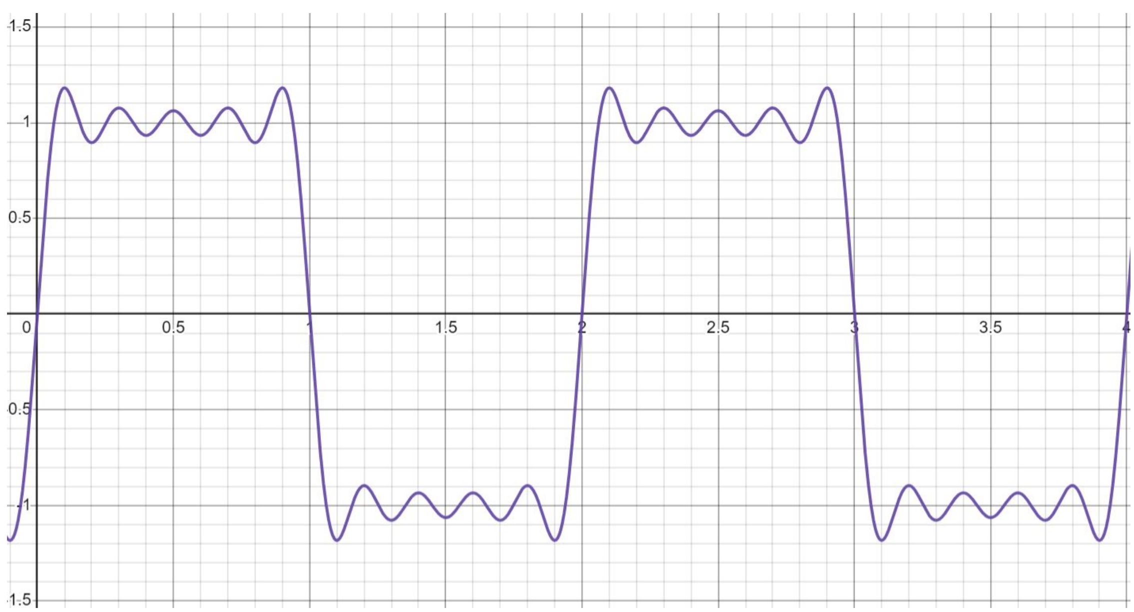


Figure 1.19: Fundamental, Third, Fifth, Seventh, & Ninth Harmonics Synthesis

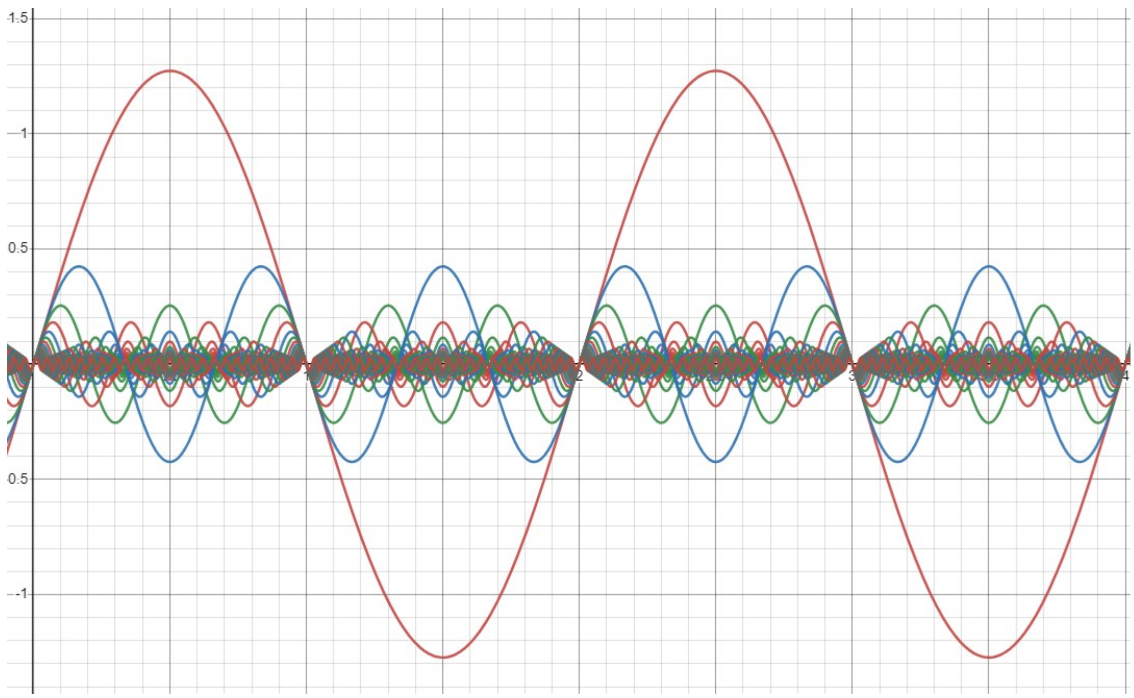


Figure 1.20: Fundamental with odd harmonics up to harmonic 49

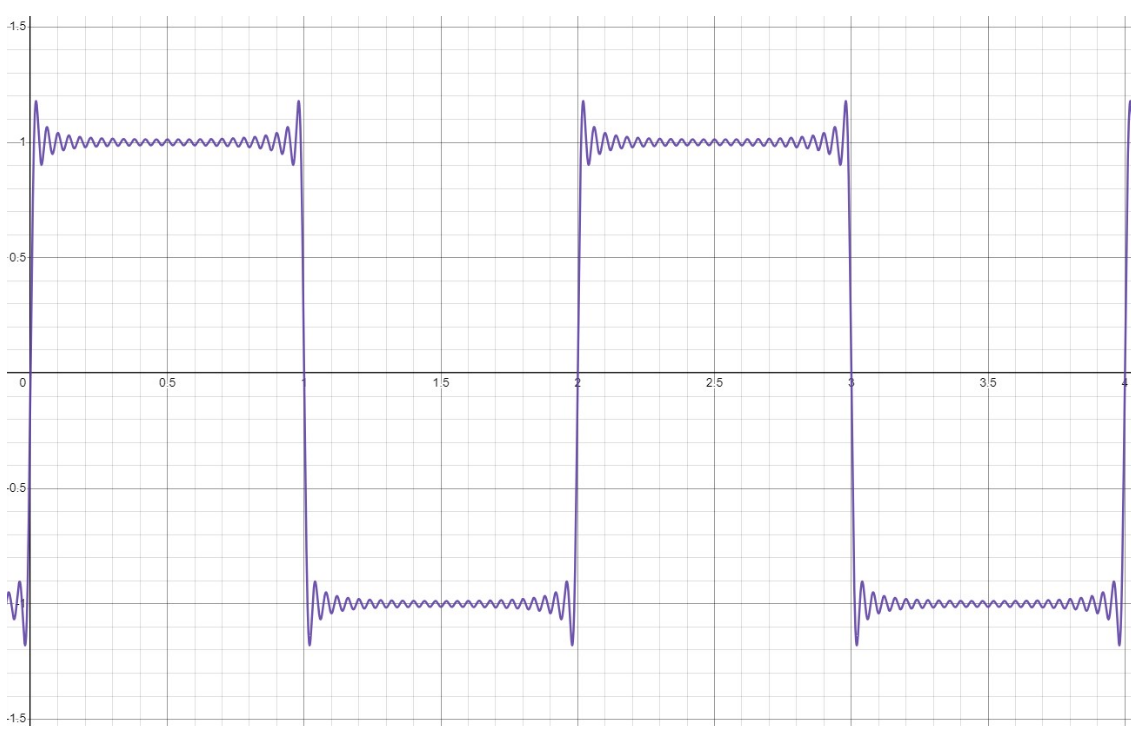


Figure 1.21: Fundamental with odd harmonics up to harmonic 49 Synthesis

1.8.3 Harmonic Analysis

Harmonic Analysis could be thought of as the opposite of, or inversely related to, Frequency Synthesis.

Harmonic Analysis involves breaking down a complex waveform into its individual sinusoidal components or harmonics and is typically achieved through techniques like Fourier Analysis.

Fourier Analysis is a mathematical technique used to decompose a complex waveform into its individual sinusoidal components, representing different frequencies.

1.8.4 Fast Fourier Transformation FFT

The "Fast Fourier Transform" (FFT) is an important measurement method in the science of audio and acoustics measurement. It converts a signal into individual spectral components and thereby provides frequency information about the signal. FFTs are used for fault analysis, quality control, and condition monitoring of machines or systems. NTI Audio [9]

Strictly speaking, the FFT is an optimized algorithm for the implementation of the "Discrete Fourier Transformation" (DFT). A signal is sampled over a period of time and divided into its frequency components. These components are single sinusoidal oscillations at distinct frequencies each with their own amplitude and phase. This transformation is illustrated in figure 1.22 FFT Time, Frequency, Amplitude - Image from NTI Audio [9]. Over the time period measured, the signal contains 3 distinct dominant frequencies. NTI Audio [9].

There are a variety of uses that can benefit from viewing the frequency spectrum of a signal. Using the FFT math function on a time domain signal provides the user with frequency domain information and can provide the user a different view of the signal quality, resulting in improved measurement productivity when troubleshooting a device-under-test. Tektronix[10]

Examples include:

- Analyze harmonics in power lines
- Measure harmonic content and distortion in systems
- Characterize noise in DC power supplies
- Test impulse response of filters and systems
- Analyze vibration Tektronix[10]

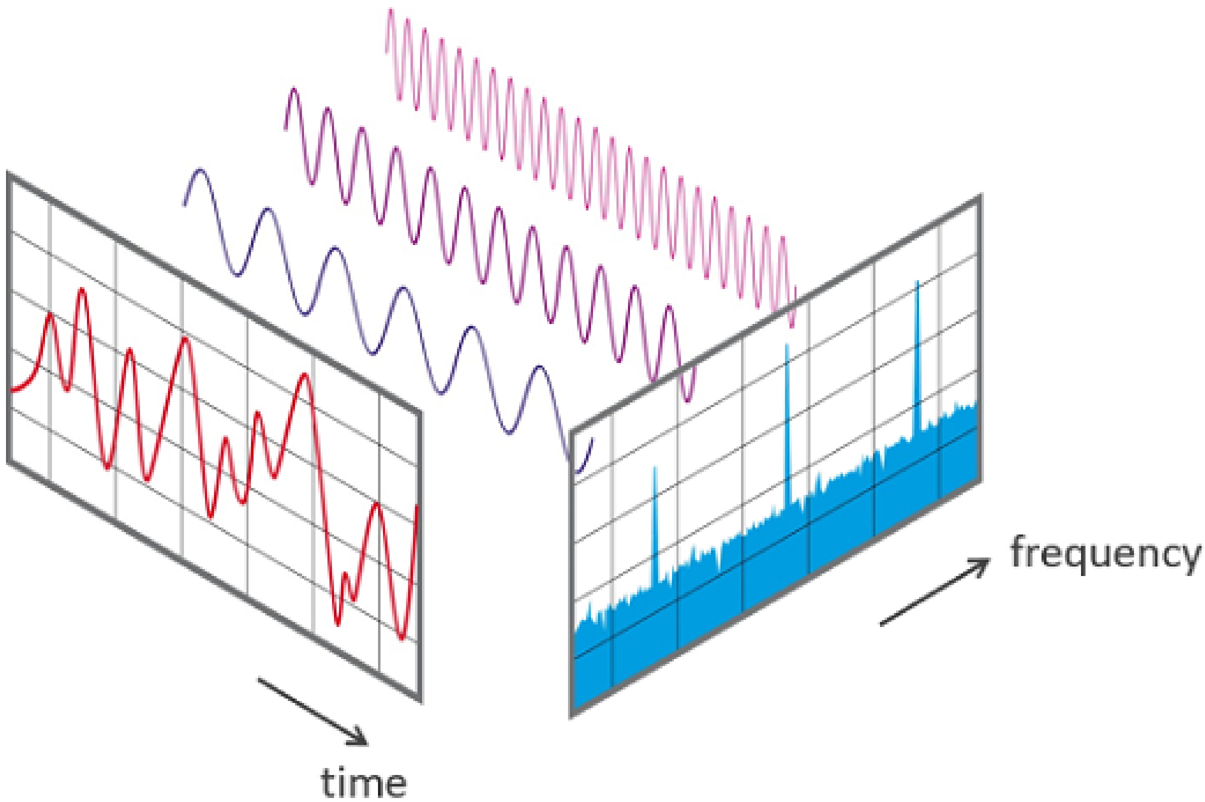


Figure 1.22: FFT Time, Frequency, Amplitude - Image from NTI Audio [9]

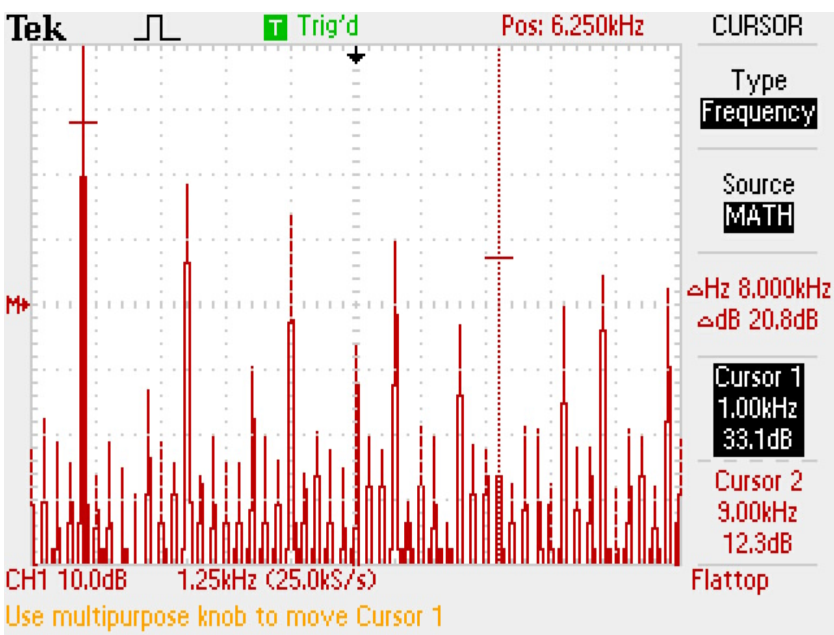


Figure 1.23: Oscilloscope FFT Measurement

Week 2

Pulse Theory and RC Circuits

2.1 Objectives:

Understand Pulse Waveform Terminology and Formulas.

- Develop a solid understanding of pulse waveform terminology, including Period, Pulse Width, Pulse Space, Pulse Repetition Frequency, Duty Cycle, Tilt, Rise Time, Average Pulse Amplitude, Average Waveform Voltage, Capacitor Charge, Cycles to Stabilization, V_{Max}, and V_{Min}.

Calculate Critical Frequencies using Tilt and Rise Time.

- Develop the ability to calculate critical frequencies using tilt and rise time measurements. Understand how series and parallel capacitance influence these parameters and the critical frequencies.

Identify Capacitor Charge Percentage in Terms of Time and τ.

- Students should be able to identify and calculate capacitor charge percentages concerning time and the time constant (τ) in RC circuits.

Calculate and Predict Waveforms for RC Integration Circuits

- Develop a comprehensive understanding of RC Integration circuits. Gain proficiency in calculating and predicting output waveforms for these circuits.

Calculate and Predict Waveforms for RC Differentiation Circuits

- Develop a comprehensive understanding of RC Differentiation circuits. Gain proficiency in calculating and predicting output waveforms for these circuits.

By achieving these objectives, students will acquire a deep understanding of waveform characteristics, pulse waveform terminology, frequency synthesis, critical frequencies, capacitor charge calculations, and the principles behind RC integration, differentiation, and sine wave analysis. These objectives aim to enhance their knowledge and proficiency in working with various waveforms and circuits.

2.2 References:

- Solid State Pulse Circuits [11]

2.3 Pulse Waveform Characteristics

2.3.1 Ideal Pulse Waveform

The **Ideal Pulse Waveform** has perfectly vertical leading and lagging edges (instantaneous rise and fall times) and perfectly flat tops and bottoms.

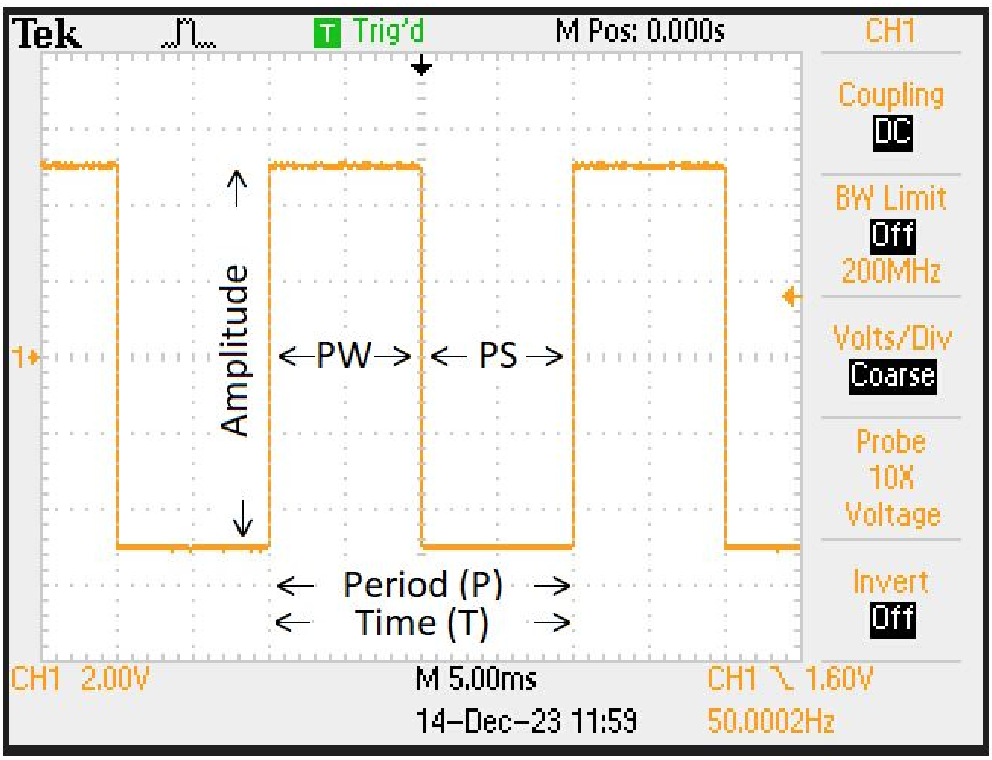


Figure 2.1: Ideal Pulse Waveform

2.3.2 Pulse Waveform Terminology:

- **Period (P):** The time it takes for one complete cycle, from one rising edge to the next rising edge.
- **Pulse Width (PW):** Also known as Time High/On, it is the duration of time during which the waveform is at its maximum amplitude. Typically measured at 50% of the waveform amplitude.
- **Pulse Space (PS):** Also known as Time Low/Off, it is the duration of time during which the waveform is at its minimum amplitude. Also measured at 50% of the waveform amplitude.
- **Pulse Repetition Frequency (PRF):** The reciprocal of the period, representing the number of pulses per unit time.

$$PRF = \frac{1}{Period}$$

- **Duty Cycle (DC%):** The ratio of the pulse width to the period, expressed as a percentage.

$$DC\% = \frac{PW}{Period} \times 100$$

2.3.3 Practical Pulse Waveform

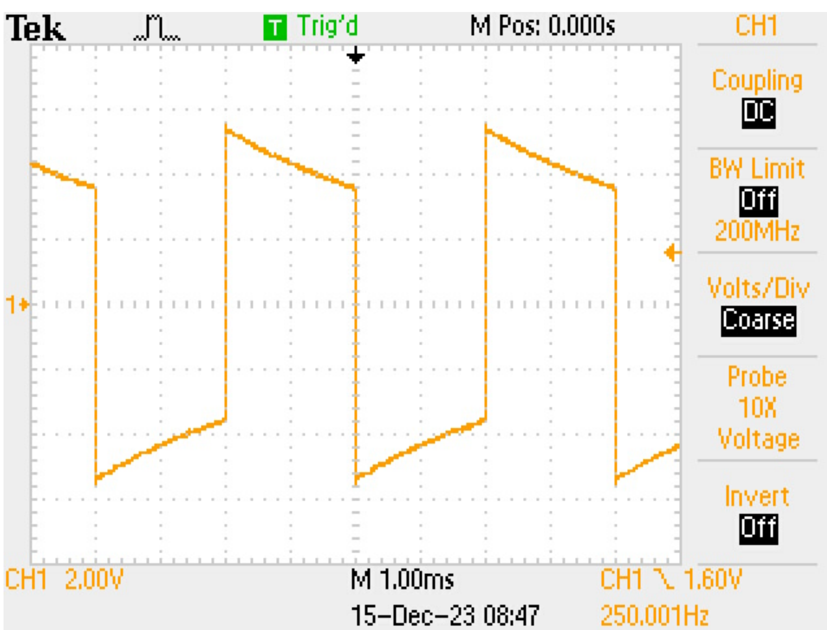


Figure 2.2: Practical Pulse Waveform

The RC circuit practical pulse waveform as seen in Fig 2.2 at 250Hz is outputting a square wave with significant tilt. Tilt represents circuit low-frequency attenuation.

2.3.4 Vmax & Vmin

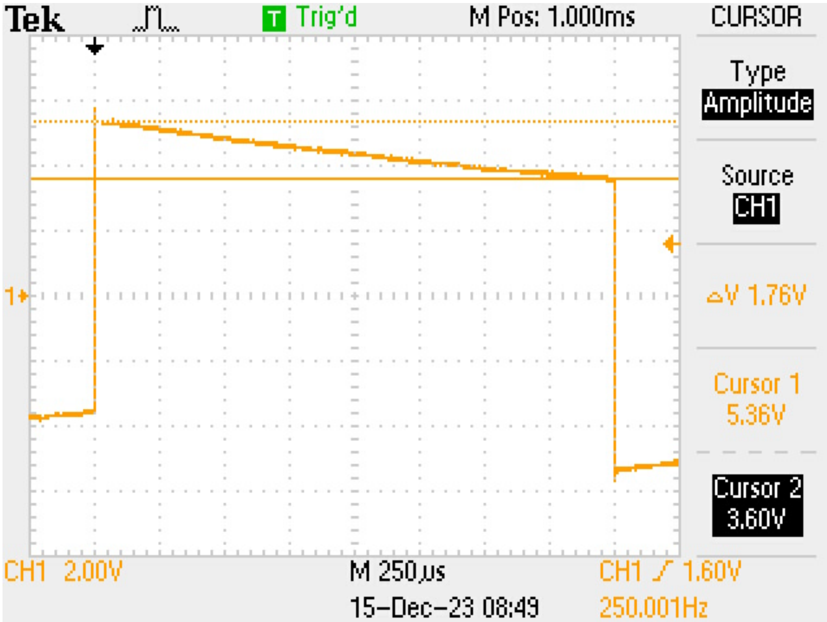


Figure 2.3: Measuring Vmax & Vmin on a Tilt waveform

Defined

- **Vmax** is the highest point of the positive peak.
- **Vmin** is the lowest point of the positive peak.

Measurement Steps

1. Lower the square-wave test signal frequency until tilt is observable.
2. Expand the waveform across the display of the oscilloscope using the horizontal control, see Fig 2.3.
3. Use the cursors to measure Vmax and Vmin of the pulse, see Fig 2.3. Cursor 1 (Vmax) is 5.56V and Cursor 2 (Vmin) is 3.6V.

2.3.5 Rise & Fall Time

Defined

- Rise Time (t_r) is defined as the time required for the voltage to go from 10% to 90% of the average pulse amplitude (APA).

  $APA = \frac{Vmax+Vmin}{2}$

- Fall Time (t_f is defined as the time required for the voltage to go from 90% to 10% of the average pulse amplitude (APA).

  $APA = \frac{Vmax+Vmin}{2}$

Measurements Steps

1. Increase the square-wave test signal frequency until the tilt is mostly gone, see Fig 2.4.
2. Use the **Fine** Volts/Div to adjust the signal to have exactly 5 major divisions peak to peak, see Fig 2.4.
3. Center the waveform vertically, trigger on the rising edge, and set the trigger point on the y-axis.
4. Zoom in on the rising edge of the waveform, see Fig 2.5
5. Measure the rise time at the -2 divisions (10%) and +2 divisions (90%), see Fig 2.5.

   $t_r = 410nS$

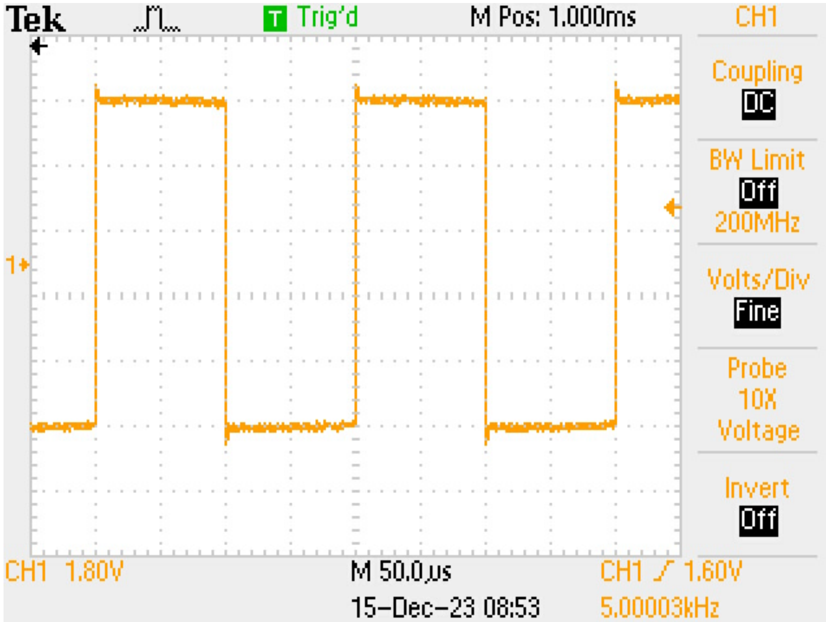


Figure 2.4: Rise Time Set-up Waveform

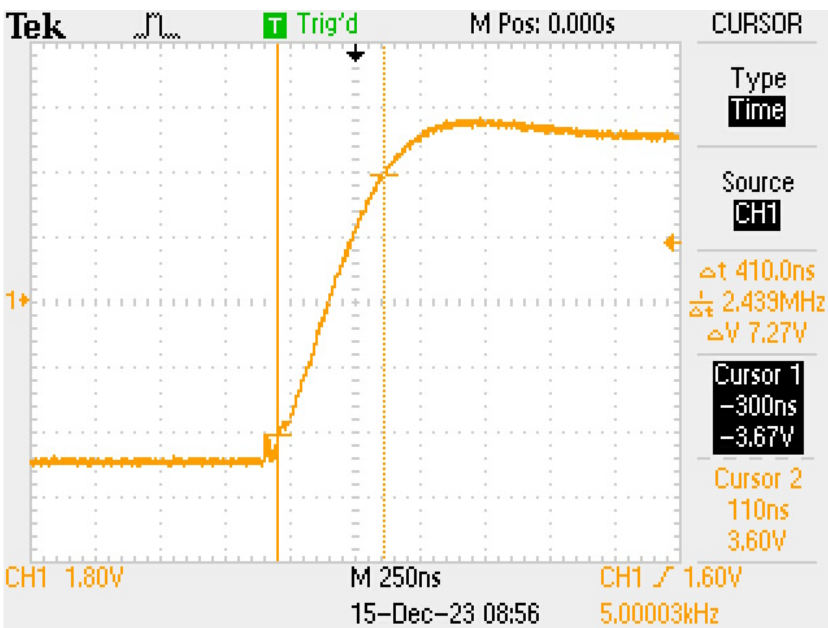


Figure 2.5: Rise Time Measurement

2.3.6 APA (Average Pulse Amplitude) formula

$$APA = \frac{Vmax + Vmin}{2}$$

2.3.7 Tilt formulas

$$Tilt = \frac{Vmax - Vmin}{APA}$$

$$Tilt\% = \frac{Vmax - Vmin}{APA} \times 100$$

2.3.8 AWV (Average Waveform Voltage)

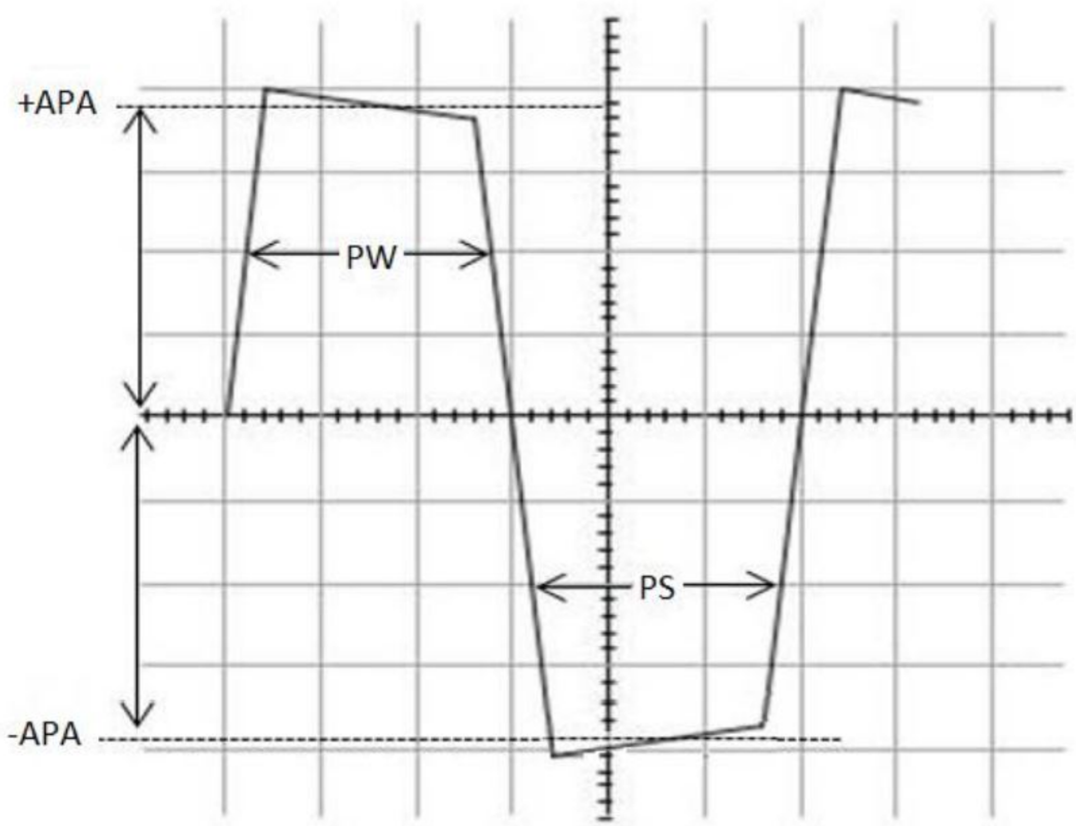


Figure 2.6: AWV (Average Waveform Voltage)

AWV formula

$$AWV = \frac{(+APA \times PW) + (-APA \times PS)}{Period}$$

AWV explained

If the positive peak and negative peak are equal in amplitude, and the pulse width and pulse space are equal, then the waveform is symmetric, and the average voltage is indeed zero. This is because, over a complete cycle, the positive and negative voltages cancel each other out.

However, if the positive peak and negative peak amplitudes are different, or if the pulse width and pulse space are different, the waveform is no longer symmetric, and the average voltage will not be zero. In this case, you would use the above formula to calculate the average voltage.

AWV examples

1. AWV Practice Problem

   Given:

   - $+APA = 8V$
   - $-APA = -10V$
   - $PW\&PS = 50\mu S$

$$AWV = \frac{(+APA \times PW) + (-APA \times PS)}{Period}$$

$$AWV = \frac{(8V \times 50\mu S) + (-10V \times 50\mu S)}{100\mu S}$$

$$AWV = \frac{400 \times 10^{-6} + -500 \times 10^{-6}}{100 \times 10^{-6}}$$

$$AWV = \frac{-100 \times 10^{-6}}{100 \times 10^{-6}}$$

$$AWV = -1V$$

2. AWV Practice Problem

   Given:

   - $+APA = 8V$
   - $-APA = -10V$
   - $PW = 30\mu S$
   - $PS = 10\mu S$

$$AWV = \frac{(+APA \times PW) + (-APA \times PS)}{Period}$$

$$AWV = \frac{(8V \times 30\mu S) + (-10V \times 10\mu S)}{40\mu S}$$

$$AWV = \frac{240 \times 10^{-6} + -100 \times 10^{-6}}{40 \times 10^{-6}}$$

$$AWV = \frac{140 \times 10^{-6}}{40 \times 10^{-6}}$$

$$AWV = 3.5V$$

2.4 RC Circuit 1 Review

2.4.1 DC Analysis

VR1 and VRgen:

When DC power is applied to the circuit, C1 will charge to the voltage potential determined by VR3 plus VR4. Once charged there will be no DC current in the R1/Rgen branch, therefore VR1 and Rgen will each equal $0V_{DC}$.

- $VRgen_{DC} = 0VDC$
- $VR1_{DC} = 0VDC$

VR2:

- $VR2_{DC} = IR2 \times R2$
- $VR2_{DC} = \frac{VCC}{R2+R3+R4} \times R2$

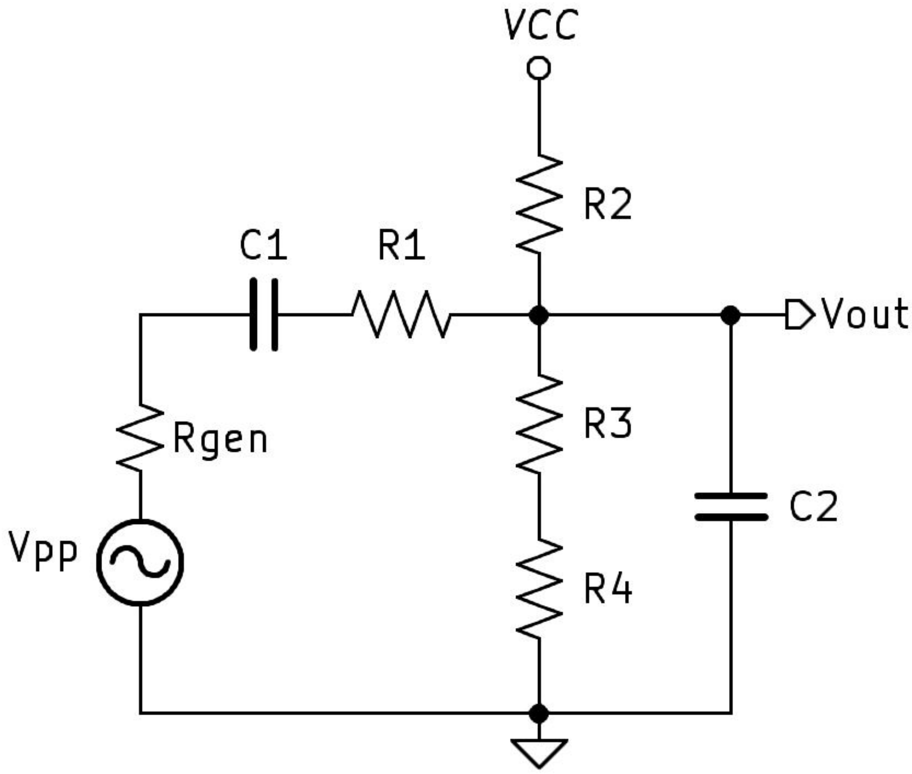


Figure 2.7: RC Circuit 1

VR3:

- $VR3_{DC} = IR3 \times R3$
- $VR3_{DC} = \frac{VCC}{R2+R3+R4} \times R3$

VR4:

- $VR4_{DC} = IR4 \times R4$
- $VR4_{DC} = \frac{VCC}{R2+R3+R4} \times R4$

Vout DC:

- $Vout_{DC} = VR3_{DC} + VR4_{DC}$

VC1 DC:

- $VC1_{DC} = VR3_{DC} + VR4_{DC}$

VC1 DC:

- $VC1_{DC} = VR3_{DC} + VR4_{DC}$

2.4.2 AC Analysis at Mid-band

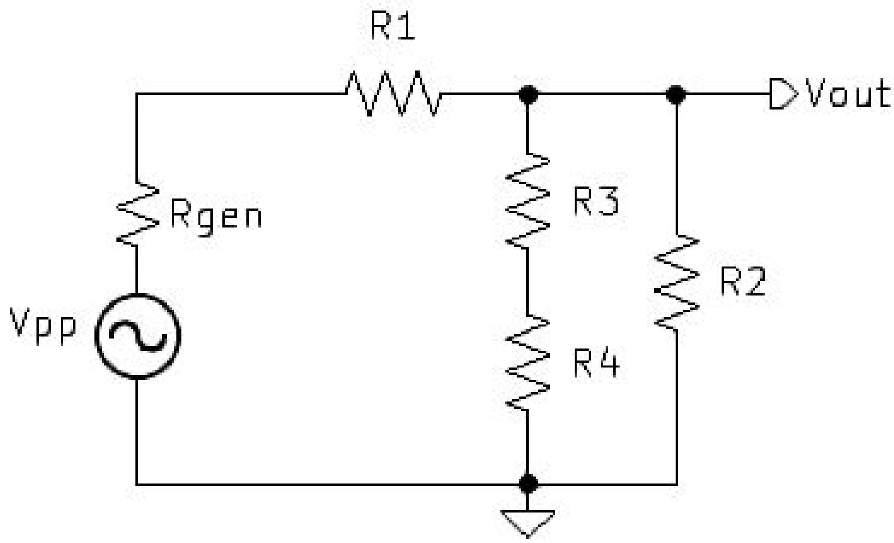


Figure 2.8: RC Circuit 1

At mid-band, the coupling capacitor C1 will act like a short to couple the AC signal to the load. Only when frequencies are nearing the upper critical frequency will C2 begin to react with the circuit, at mid-band frequencies C2 is open or high impedance relative to the circuit.

Resistance Total:

- $R_{Total} = R_{gen} + R1 + \frac{1}{\frac{1}{R3+R4}+\frac{1}{R2}}$

Peak to Peak Current Total:

- $I_{Total} = \frac{Vpp}{R_{Total}}$

VRgen:

- $VR_{gen} = I_{Total} \times R_{gen}$

VR1:

- $VR1 = I_{Total} \times R1$

VR2:

- $VR2 = I_{Total} \times \frac{1}{\frac{1}{R3+R4}+\frac{1}{R2}}$

VR3 and VR4:

- $VR3 = \frac{VR2}{R3+R4} \times R3$
- $VR4 = \frac{VR2}{R3+R4} \times R4$

2.4.3 Frequency Critical Low

Coupling or series capacitance will affect FC_{low}.

Derive the Critical Frequency Formula:

- $X_C = \frac{1}{2\pi FC}$
- At Critical Frquency, $X_C = R_{Thev}$
- $R_{Thev} = \frac{1}{2\pi F_C C}$
- $F_C = \frac{1}{2\pi R_{Thev} C}$

Thevenin Resistance C1:

Below mid-band frequencies, the bypass capacitor C2 will act like an open.

- $R_{ThevC1} = R1 + \frac{1}{\frac{1}{R3+R4}+\frac{1}{R2}} + R_{gen}$

Thevenin Resistance C2:

At and above mid-band frequencies, the coupling capacitor C1 will act like a short.

- $R_{ThevC2} = \frac{1}{\frac{1}{R3+R4}+\frac{1}{R2}+\frac{1}{R1+R_{gen}}}$

2.5 Waveform Distortion

Consider an amplifier or filter circuit with a particular set of band-pass frequencies. What will happen if a square wave is applied to a circuit that limits the amplitude of the square wave harmonic content of the frequencies outside the pass-band? The answer is **waveform distortion**. The two types of square wave distortion that are useful for circuit frequency response analysis are Tilt and Rise Time (t_r).

- **Tilt** represents low frequency amplitude attenuation and can be used to determine Frequency Critical Low (FC_L).
- **Rise Time (t_r)** represents high-frequency amplitude attenuation and can be used to determine Frequency Critical High (FC_H).

2.6 Tilt and Frequency Critical Low (FC_L)

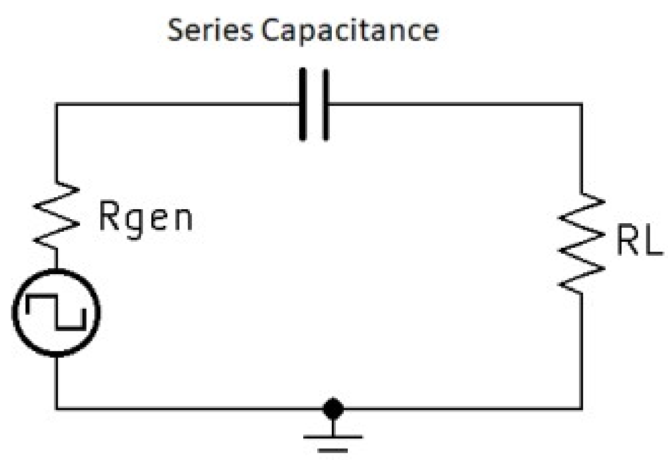


Figure 2.9: Series Capacitance, Tilt, & FC_L

Consider the circuit of figure 2.9 Series Capacitance, Tilt, & FC_L.

- Series capacitance will allow high frequencies through but will attenuate frequencies below FC_L.
- The horizontal component of the square wave represents higher to lower (left to right) frequency amplitude of the odd harmonics.
- X_C of the series capacitance increases as frequency decreases.
- Tilt represents the attenuation of low frequencies.

2.6.1 FC_{Low} formula using Tilt

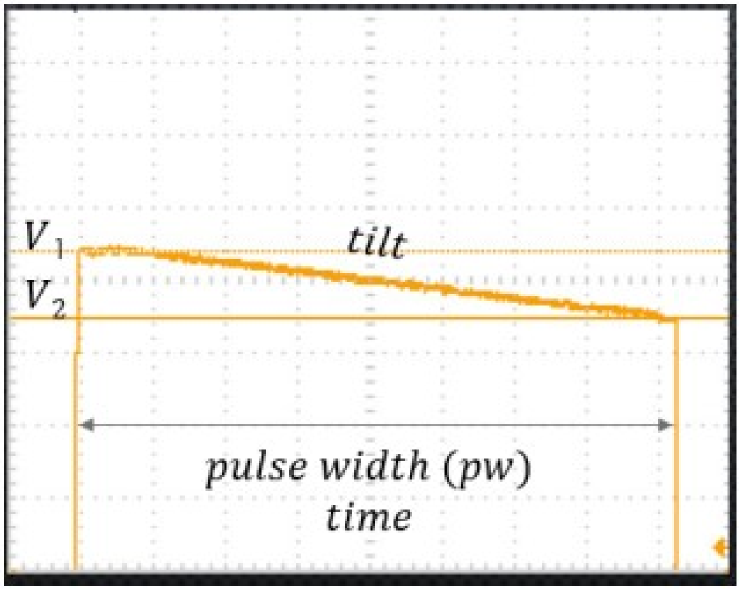


Figure 2.10: Tilt Measurement

$$FC_{Low} = \frac{fractional\ tilt}{2\pi PW}$$

$$fractional\ tilt = \frac{V1 - V2}{APA}$$

$$APA = \frac{V1 + V2}{2}$$

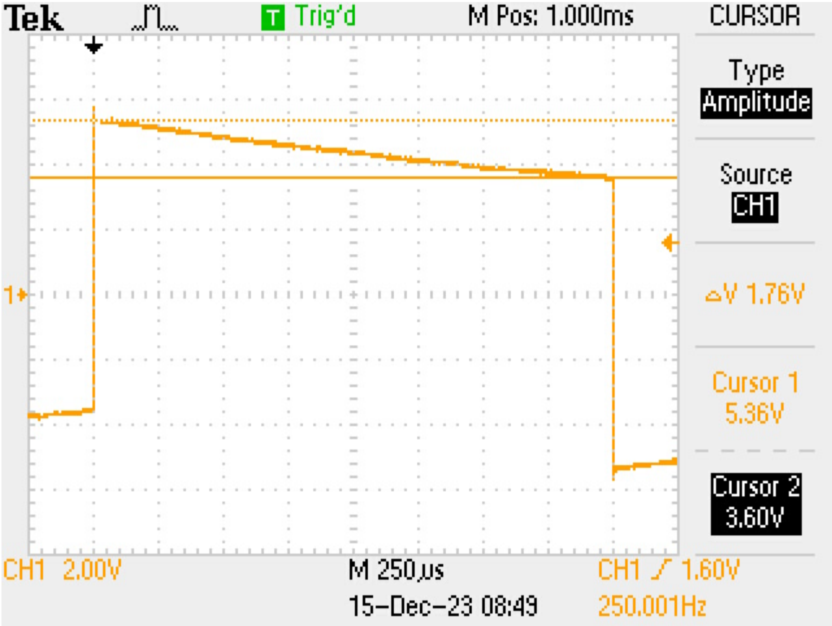


Figure 2.11: Measuring Vmax & Vmin on a Tilt waveform

2.6.2 FC_{Low} Tilt Example

Example 1. See Figure 2.11 Measuring Vmax & Vmin on a Tilt waveform.

- $V_{Cursor1} = 5.36V$
- $V_{Cursor2} = 3.6V$
- $Frequency = 250Hz$

Solution Steps:

1. Find APA:

$$APA = \frac{V1 + V2}{2}$$

$$APA = \frac{5.36V + 3.6V}{2}$$

$$APA = \frac{8.96V}{2}$$

$$\boxed{APA = 4.48V}$$

2. Find Fractional Tilt:

$$fractional\ tilt = \frac{V1 - V2}{APA}$$

$$fractional\ tilt = \frac{5.36V - 3.6V}{4.48V}$$

$$fractional\ tilt = \frac{1.76V}{4.48V}$$

$$\boxed{fractional\ tilt = 0.393}$$

3. Find Pulse Width:

$$PW = \frac{Period}{2}$$

$$Period = \frac{1}{Frequency}$$

$$Period = \frac{1}{250HZ}$$

$$Period = 4mS$$

$$PW = \frac{4mS}{2}$$

$$\boxed{PW = \ 2mS}$$

4. Find Frequency Critical Low:

$$FC_{Low} = \frac{fractional\ tilt}{2\pi PW}$$

$$FC_{Low} = \frac{0.393}{2\pi(2mS)}$$

$$FC_{Low} = \frac{0.393}{12.566 \times 10^{-3}}$$

$$\boxed{FC_{Low} = 31.274HZ}$$

2.7 Rise Time t_r and Frequency Critical High FC_H

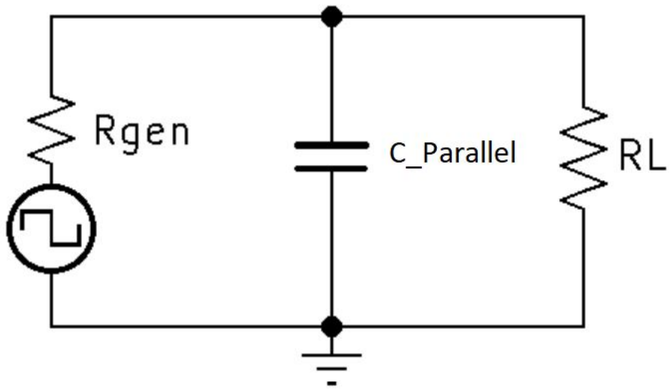


Figure 2.12: Parallel Capacitance, Rise Time, & FC_H

Consider the circuit of figure 2.12 Parallel Capacitance, Rise Time, & FC_H.

- **High Frequency Distortion** is caused by **parallel capacitance**.
- **Parallel capacitance** can include stray capacitance, probe capacitance, generator capacitance, and device capacitance.
- The attenuation of high frequencies caused by parallel capacitance will slow or decrease the **Rise Time** t_r of a measured square wave.

2.7.1 FC_{High} formula using Rise Time t_r

Using Pulse Theory or Square-Wave Analysis, we can calculate the circuit's high critical frequency using the following formula:

$$FC_{High} = \frac{0.35}{t_r}$$

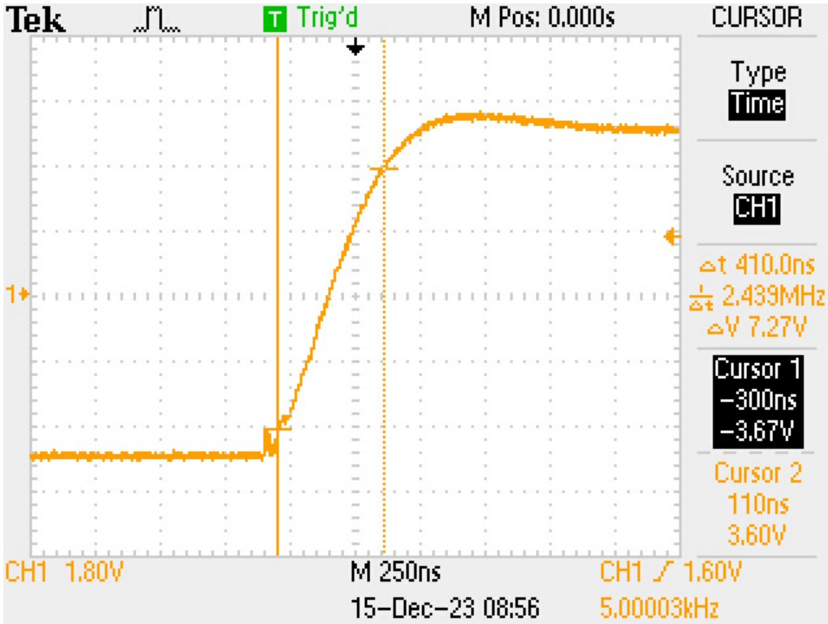


Figure 2.13: Rise Time Measurement

2.7.2 FC_{High} and Rise Time t_r Example

See Figure 2.13 Rise Time Measurement.

1. Measure Rise Time

$$t_r = 410nS$$

2. Calculated FC_{High}

$$FC_{High} = \frac{0.35}{t_r}$$

$$FC_{High} = \frac{0.35}{410nS}$$

$$\boxed{FC_{High} = 853.659Khz}$$

2.8 Capacitor Charge Formula

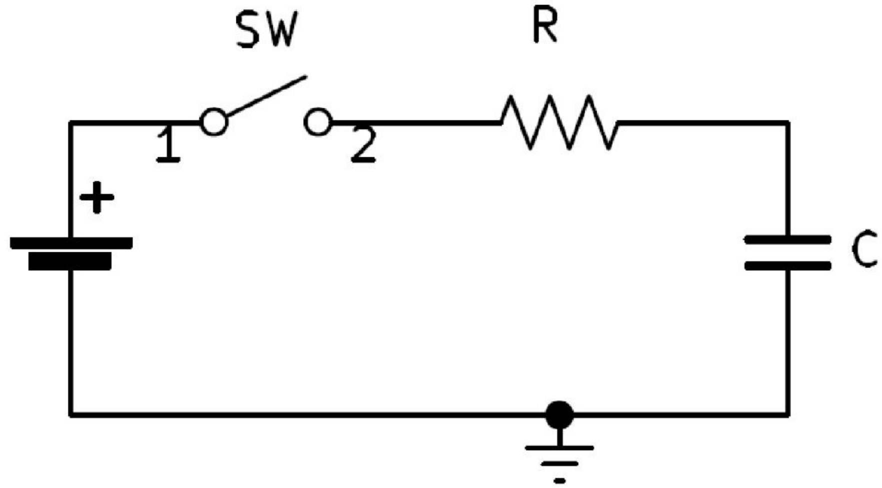


Figure 2.14: Capacitor Charge

After the switch is closed, how long will it take for the capacitor to fully charge?

$$V_{Capacitor} = V_{Final} - (V_{Final} - V_{Initial})e^{(\frac{-t}{\tau})}$$

* V_{Final} is the voltage that the capacitor would charge or discharge to if time is greater than 5τ.
* $V_{Initial}$ is the initial voltage that the capacitor is initially at.
* t is the amount of charge time.
* Tau (τ) is RC.

2.8.1 Capacitor Charge Formula in terms of % and time vs. τ

Percent of Charge at time equals 1τ:

$$V_C = 100\%V_{Fin} - (100\%V_{Fin} - 0V_{init})e^{(\frac{-1}{1})}$$

* $V_C = 100\%V_{Fin} - (100\%V_{Fin})e^{-1}$
* $V_C = 100\%V_{Fin} - (100\%V_{Fin})(0.36787944)$
* $V_C = 100\%V_{Fin} - 36.787944\%V_{Fin}$

* $V_C = 63.212056\% V_{Fin}$

✓ At 1τ, the capacitor will charge to 63.212% of the final voltage.

Percent of Charge at time equals 2τ:

$$V_C = 100\% V_{Fin} - (100\% V_{Fin} - 0 V_{init}) e^{(\frac{-2}{1})}$$

* $V_C = 100\% V_{Fin} - (100\% V_{Fin}) e^{-2}$
* $V_C = 100\% V_{Fin} - (100\% V_{Fin})(0.13533528)$
* $V_C = 100\% V_{Fin} - 13.533528\% V_{Fin}$
* $V_C = 86.466471\% V_{Fin}$

✓ At 2τ, the capacitor will charge to 86.466% of the final voltage.

Percent of Charge at time equals 3τ:

$$V_C = 100\% V_{Fin} - (100\% V_{Fin} - 0 V_{init}) e^{(\frac{-3}{1})}$$

* $V_C = 100\% V_{Fin} - (100\% V_{Fin}) e^{-3}$
* $V_C = 100\% V_{Fin} - (100\% V_{Fin})(0.049787068)$
* $V_C = 100\% V_{Fin} - 4.9787068\% V_{Fin}$
* $V_C = 95.02129316\% V_{Fin}$

✓ At 3τ, the capacitor will charge to 95.021% of the final voltage.

Percent of Charge at time equals 4τ:

$$V_C = 100\% V_{Fin} - (100\% V_{Fin} - 0 V_{init}) e^{(\frac{-4}{1})}$$

* $V_C = 100\% V_{Fin} - (100\% V_{Fin}) e^{-4}$

* $V_C = 100\%V_{Fin} - (100\%V_{Fin})(0.018315639)$
* $V_C = 100\%V_{Fin} - 1.8315639\%V_{Fin}$
* $V_C = 98.1684361\%V_{Fin}$

✓ At 4τ, the capacitor will charge to 98.168% of the final voltage.

Percent of Charge at time equals 5τ:

$$V_C = 100\%V_{Fin} - (100\%V_{Fin} - 0V_{init})e^{(\frac{-5}{1})}$$

* $V_C = 100\%V_{Fin} - (100\%V_{Fin})e^{-5}$
* $V_C = 100\%V_{Fin} - (100\%V_{Fin})(0.006737947)$
* $V_C = 100\%V_{Fin} - 0.6737947\%V_{Fin}$
* $V_C = 99.3262\%V_{Fin}$

✓ At 5τ, the capacitor will charge to 99.326% of the final voltage.

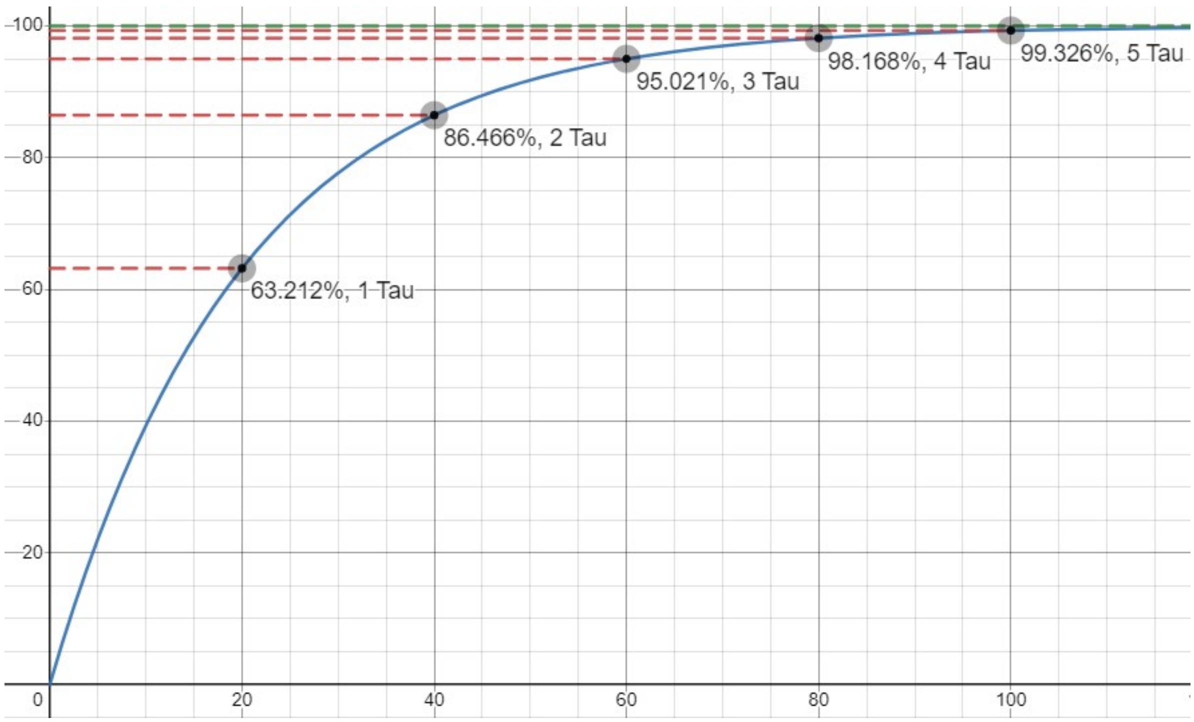


Figure 2.15: Capacitor % of Charge vs. Tau

Table 2.1: RC Circuits Capacitor % of Charge vs. Tau

Tau	**VC % of Charge**
1	**63.212%**
2	**86.466%**
3	**95.021%**
4	**98.168%**
5	**99.326%**

2.9 RC Circuits Stability, Vmax & Vmin Calculations

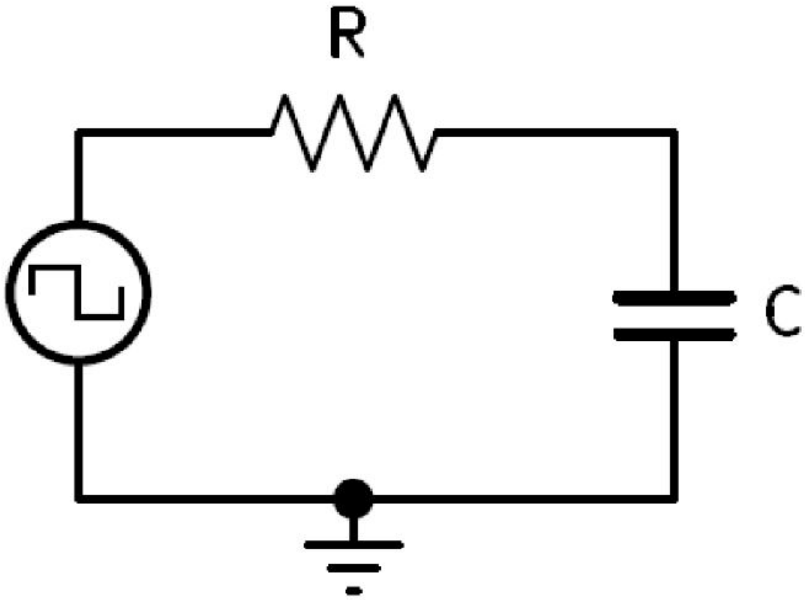


Figure 2.16: RC Circuit Stability

Consider the circuit of Figure 2.16 RC Circuit Stability. What happens to the capacitor voltage if the generator pulse width and pulse space are less than 5τ? In other words, there is not enough time to fully charge or fully discharge the capacitor. If the time on/off is less than 5 tau the capacitor voltage will not be able to charge up to the full generator voltage, likewise, Vc will not be able to fully discharge during the time off. Assuming the Duty Cycle is 50%, time on & off time are equal, the Vc waveform will appear to center itself vertically between the peak pulse width voltage and the peak pulse space voltage, as though superimposed on an average DC voltage.

$$V_{Capacitor} = V_{Final} - (V_{Final} - V_{Initial})e^{(\frac{-t}{\tau})}$$

The $V_{Capacitor}$ charge formula could be used to calculate cycle after cycle until the charge and discharge voltages stabilized. Next the number of cycles could be counted to find Cycles to Stabilization. This method can be tedious and mathematically error-prone if there are many cycles to stabilization.

2.9.1 Cycles to Stabilization Formula

Stability = 5τ (Stable?)

How many cycles to Stabilization?

If:

1 Cycle = Period

Stability = 5τ

Then:

$$\frac{Stability}{5\tau(Sec)} = \frac{1Cycle}{Period(Sec)}$$

$$Stability = \frac{1Cycle \times 5\tau(Sec)}{Period(Sec)}$$

Notice that the seconds/time unit cancels and we are left with cycles.

✓ $$\boxed{Stability_{Cycles} = \frac{5\tau}{Period}}$$

The $Stability_{Cycles}$ formula only works if the Duty Cycle is 50%. If the generator signal is not at 50% Duty Cycle, the capacitor charge formula long method must be used to calculate the number of cycles to stability.

2.9.2 Vmax and Vmin formulas

Now that we can predict the number of cycles to stabilization,What will the stable capacitor voltage waveform voltage be? What will the maximum and minimum voltage be?

$$V_C = V_{fin} - (V_{fin} - V_{In})e^{\frac{-t}{RC}}$$
$$t = PW$$
$$V_C = V_{Max}$$
$$V_{gen+} = V_{fin}$$
$$V_{In} = V_{Min}$$
$$V_{Min} = V_{gen+} - V_{Max}$$
$$V_{In} = V_{gen+} - V_{Max}$$
$$V_{Max} = V_{gen+} - (V_{gen+} - (V_{gen+} - V_{Max}))e^{\frac{-PW}{RC}}$$
$$V_{Max} = V_{gen+} - (V_{gen+} - V_{gen+} + V_{Max})e^{\frac{-PW}{RC}}$$
$$V_{Max} = V_{gen+} - (V_{Max})e^{\frac{-PW}{RC}}$$
$$V_{Max} + (V_{Max})e^{\frac{-PW}{RC}} = V_{gen+}$$
$$V_{Max}(1 + e^{\frac{-PW}{RC}}) = V_{gen+}$$
$$\boxed{V_{Max} = \frac{V_{gen+}}{1 + e^{\frac{-PW}{RC}}}}$$

$$\boxed{V_{Min} = V_{gen+} - V_{Max}}$$

The V_{Max} & V_{Min} formulas only works if the Duty Cycle is 50% .

2.9.3 RC Circuit Stability Example Problem:

Refer to Figure 2.16 RC Circuit Stability schematic.

Given:

- V_{gen} is 0 to 10v, 1Khz, 50%DC, square-wave.
- $R = 1K\Omega$
- $C = 1\mu F$
- $\tau = RC = 1mS$
- $5\tau = 5 \times 10^{-3} seconds\ (5mS)$
- $Period = \frac{1}{1Khz} = 1mS$
- $PW\ and\ PS = \frac{Period}{2} = 0.5mS$

Solving using the Capacitor Charge Formula:

$$V_{Capacitor} = V_{Final} - (V_{Final} - V_{Initial})e^{(\frac{-t}{\tau})}$$

- $time = 0\ to\ 0.5mSec$

$$V_C = 10V - (10V - 0V)e^{(\frac{-0.5mS}{1mS})}$$
$$V_C = 3.935V$$

- $time = 0.5\ to\ 1.0mSec$

$$V_C = 0V - (0V - 3.935V)e^{(\frac{-0.5mS}{1mS})}$$
$$V_C = 2.387V$$

- $time = 1\ to\ 1.5mSec$

$$V_C = 10V - (10V - 2.387V)e^{(\frac{-0.5mS}{1mS})}$$
$$V_C = 5.382V$$

- $time = 1.5\ to\ 2.0mSec$

$$V_C = 0V - (0V - 5.382V)e^{(\frac{-0.5mS}{1mS})}$$
$$V_C = 3.265V$$

- $time = 2.0\ to\ 2.5mSec$

$$V_C = 10V - (10V - 3.265V)e^{(\frac{-0.5mS}{1mS})}$$
$$V_C = 5.915V$$

- $time = 2.5\ to\ 3.0mSec$

$$V_C = 0V - (0V - 5.915V)e^{(\frac{-0.5mS}{1mS})}$$
$$V_C = 3.587V$$

- $time = 3.0\ to\ 3.5mSec$

$$V_C = 10V - (10V - 3.587V)e^{(\frac{-0.5mS}{1mS})}$$
$$V_C = 6.111V$$

- $time = 3.5\ to\ 4.0mSec$

$$V_C = 0V - (0V - 6.111V)e^{(\frac{-0.5mS}{1mS})}$$
$$V_C = 3.706V$$

- $time = 4.0\ to\ 4.5mSec$

$$V_C = 10V - (10V - 3.706V)e^{(\frac{-0.5mS}{1mS})}$$
$$V_C = 6.182V$$

- $time = 4.5\ to\ 5.0mSec$

$$V_C = 0V - (0V - 6.182V)e^{(\frac{-0.5mS}{1mS})}$$
$$V_C = 3.740V$$

- $time = 5.0\ to\ 5.5mSec$

$$V_C = 10V - (10V - 3.740V)e^{(\frac{-0.5mS}{1mS})}$$
$$V_C = 6.209V$$

- $time = 5.5\ to\ 6.0mSec$

$$V_C = 0V - (0V - 6.209V)e^{(\frac{-0.5mS}{1mS})}$$
$$V_C = 3.766V$$

- $time = 6.0\ to\ 6.5mSec$

$$V_C = 10V - (10V - 3.766V)e^{(\frac{-0.5mS}{1mS})}$$
$$V_C = 6.219V$$

- $time = 6.5\ to\ 7.0mSec$

$$V_C = 0V - (0V - 6.219V)e^{(\frac{-0.5mS}{1mS})}$$
$$V_C = 3.772V$$

- $time = 7.0 \ to \ 7.5mSec$

$$V_C = 10V - (10V - 3.772V)e^{(\frac{-0.5mS}{1mS})}$$
$$V_C = 6.223V$$

- $time = 7.5 \ to \ 8.0mSec$

$$V_C = 0V - (0V - 6.223V)e^{(\frac{-0.5mS}{1mS})}$$
$$V_C = 3.774V$$

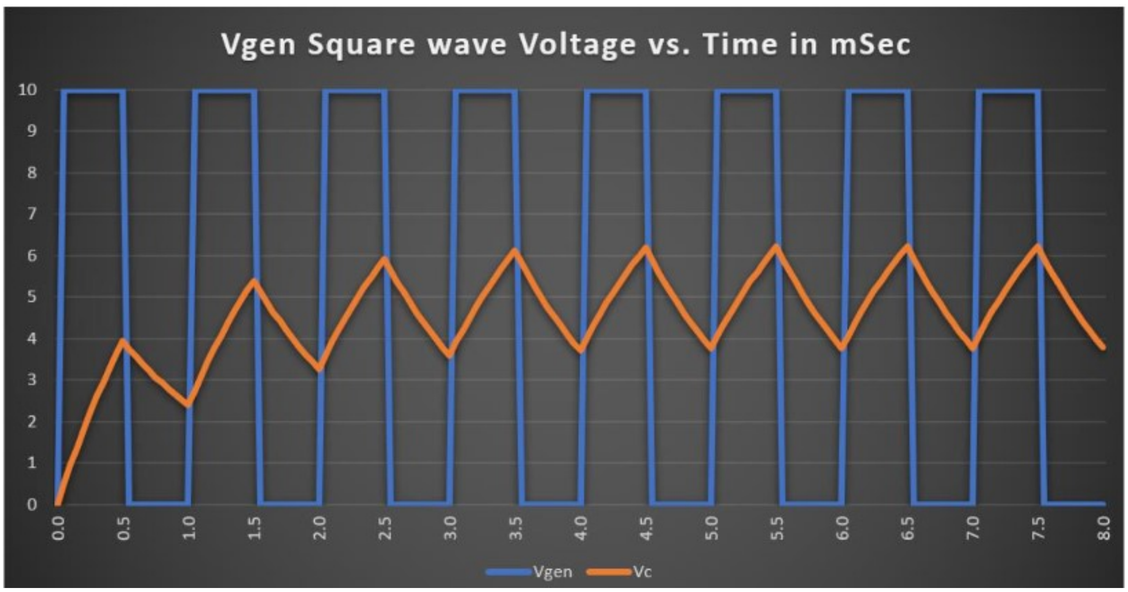


Figure 2.17: VC Stability

See Figure 2.17, observe that the capacitor waveform voltage is stable at 5τ ($5mS$) & 5 *cycles*.

Solving using the $Stability$, V_{Max}, & V_{Min} using formulas derived in Sections 2.9.1 & 2.9.2:

Refer to Figure 2.16 RC Circuit Stability schematic.

Given:

- V_{gen} is 0 to 10v, 1Khz, 50%DC, square-wave.
- $R = 1K\Omega$
- $C = 1\mu F$
- $\tau = RC = 1mS$
- $5\tau = 5 \times 10^{-3} seconds\ (5mS)$
- $Period = \frac{1}{1Khz} = 1mS$
- $PW\ and\ PS = \frac{Period}{2} = 0.5mS$

Formulas:

$$Stability_{Cycles} = \frac{5\tau}{Period}$$

$$V_{Max} = \frac{V_{gen+}}{1 + e^{\frac{-PW}{RC}}}$$

$$V_{Min} = V_{gen+} - V_{Max}$$

Calculating The Number of Cycles to Stability:

$$Stability_{Cycles} = \frac{5\tau}{Period}$$

$$Stability_{Cycles} = \frac{5(RC)}{1mS}$$

$$Stability_{Cycles} = \frac{5(1K\Omega \times 1\mu F)}{1mS}$$

$$Stability_{Cycles} = \frac{5(1mS)}{1mS}$$

$$\boxed{Stability_{Cycles} = 5\ cycles}$$

Calculating V_{Max}:

$$V_{Max} = \frac{V_{gen+}}{1 + e^{\frac{-PW}{RC}}}$$

$$V_{Max} = \frac{10V}{1 + e^{\frac{-0.5mS}{1K\Omega \times 1\mu F}}}$$

$$V_{Max} = \frac{10V}{1 + 0.606531}$$

$$\boxed{V_{Max} = 6.225vp}$$

Calculating V_{Min}:

$$V_{Min} = V_{gen+} - V_{Max}$$

$$V_{Min} = 10V - 6.225vp$$

$$\boxed{V_{Min} = 3.775vp}$$

Additionally, we can check our work by taking the average voltage of V_{Max} and V_{Min}. We know that the waveform voltage should be centered around the average generator voltage, in this case 5volts, ($5v = \frac{0+10}{2}$).

Calculating V_{avg}:

$$V_{avg} = \frac{V_{Max} + V_{Min}}{2}$$

$$V_{avg} = \frac{6.225v + 3.775v}{2}$$

$$V_{avg} = \frac{10v}{2}$$

$$\boxed{V_{avg} = 5v}$$

2.10 RC Integration

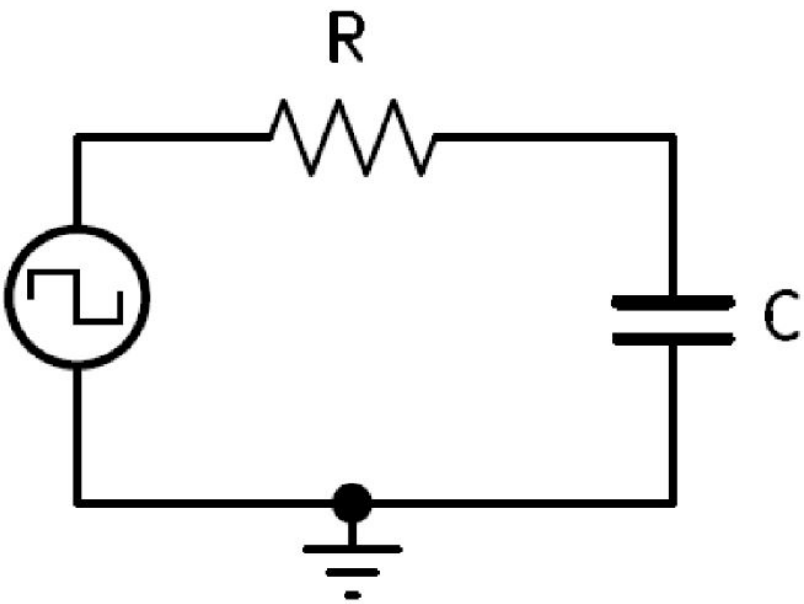


Figure 2.18: RC Integration

2.10.1 Characteristics of RC Integration Circuits

- Think Averaging!
- The output is across the capacitor
- There is not enough time for the capacitor to fully charge or discharge.
- PW and PS is less than 5τ
- $RC \geq (10 \times PW)$
- $PW(time) \leq \frac{RC}{10}$ (not enough time)
- Additional Formulas:
  * $Stability_{Cycles} = \frac{5\tau}{Period}$
  * $V_{Max} = \frac{V_{gen+}}{1+e^{\frac{-PW}{RC}}}$
  * $V_{Min} = V_{gen+} - V_{Max}$

2.10.2 RC Integration Circuit Example

Calculate an RC Integration Circuit using a $1Khz$ frequency and a $1\mu F$ capacitor.

Given:

- V_{gen} = 0vp to 10vp, 50% DC, square-wave.
- $F_{gen} = 1Khz$
- $C = 1\mu F$

Find charge/discharge time($time$):

$$Period = \frac{1}{Frequency}$$

$$Period = \frac{1}{1Khz}$$

$$Period = 1mS$$

$$time = \frac{Period}{2}$$

$$time = \frac{1mS}{2}$$

$$\boxed{time = 0.5mS}$$

Find the series resistance (R):

$$PW(time) \leq \frac{RC}{10}$$

$$0.5mS \leq \frac{R(1\mu F)}{10}$$

$$5mS \leq R(1\mu F)$$

$$\frac{5mS}{1\mu F} \leq R$$

$$R \geq \frac{5mS}{1\mu F}$$

$$\boxed{R \geq 5K\Omega}$$

Find the number of cycles to stabilization ($Stability_{Cycles}$):

$$Stability_{Cycles} = \frac{5\tau}{Period}$$

$$Stability_{Cycles} = \frac{5(5K\Omega \times 1\mu F)}{1mS}$$

$$Stability_{Cycles} = \frac{5(5mS)}{1mS}$$

$$Stability_{Cycles} = \frac{25mS}{1mS}$$

$$\boxed{Stability_{Cycles} = 25_{Cycles}}$$

Find V_{Max}:

$$V_{Max} = \frac{V_{gen+}}{1 + e^{\frac{-PW}{RC}}}$$

$$V_{Max} = \frac{10vp}{1 + e^{\frac{-0.5mS}{5K\Omega \times 1\mu F}}}$$

$$V_{Max} = \frac{10vp}{1 + e^{\frac{-0.5mS}{5K\Omega \times 1\mu F}}}$$

$$V_{Max} = \frac{10vp}{1 + 0.905}$$

$$\boxed{V_{Max} = 5.250vp}$$

Find V_{Min}:

$$V_{Min} = V_{gen+} - V_{Max}$$

$$V_{Min} = 10vp - 5.250vp$$

$$\boxed{V_{Min} = 4.750vp}$$

******Add here Measured Image Integrator Gen=10vp squarewave 1KHZ Waveform, R=5K, C=1microF******

2.11 RC Differentiation

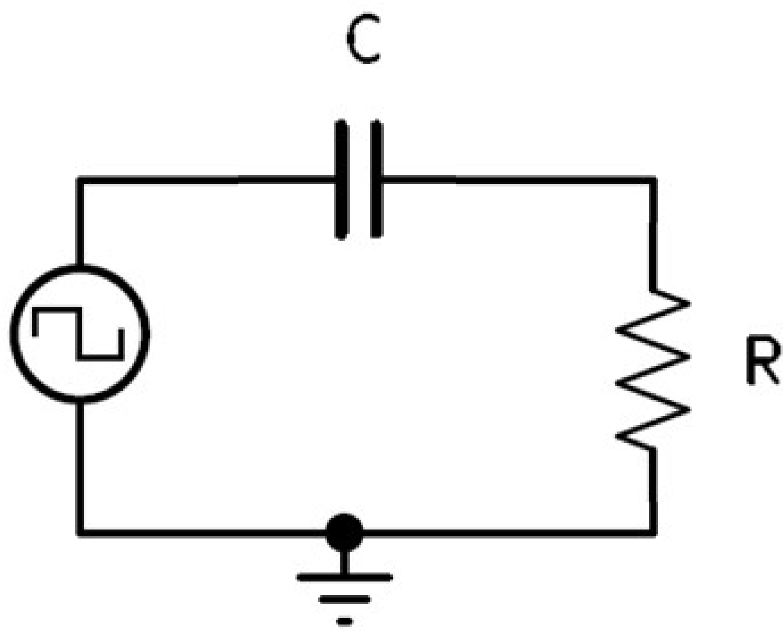


Figure 2.19: RC Differentiation

2.11.1 Characteristics of RC Differentiation Circuits

- When RC is less than one-tenth of the pulse width, the capacitor is charged very rapidly. Only a brief pulse of current is necessary to charge and discharge the capacitor at the beginning and end of the pulse. The resultant waveform of the resistor voltage is a series of positive and negative spikes at the pulse leading and lagging edges, respectively. Bell p.56[11]
- The output is across the resistor.
- There is more than enough time for the capacitor to fully charge or discharge.
- PW and PS is more than 5τ
- $RC \leq \frac{PW}{10}$
- $PW(time) \geq 10 \times RC$ (lots of time)

2.11.2 RC Differentiation Circuit Example

Calculate an RC Differentiation Circuit using a $1Khz$ frequency and a $1\mu F$ capacitor.

Given:

- V_{gen} = 0vp to 10vp, 50% DC, square-wave.
- $F_{gen} = 1Khz$
- $R = 5K\Omega$

Find charge/discharge time($time$):

$$Period = \frac{1}{Frequency}$$

$$Period = \frac{1}{1Khz}$$

$$Period = 1mS$$

$$time = \frac{Period}{2}$$

$$\boxed{time = 0.5mS}$$

Find the series capacitance (C):

$$PW(time) \geq 10 \times RC$$

$$0.5mS \geq 10 \times 5K\Omega \times C$$

$$0.5mS \geq 50K\Omega \times C$$

$$\frac{0.5mS}{50K\Omega} \geq C$$

$$\boxed{C = 10nF\ or\ 10{,}000pF}$$

Calculating the capacitor voltage based on Tau:

In Section 2.8 Capacitor Charge Formula the capacitor charge formula was simplified in terms of Tau. We can now apply this to our current example to find the capacitor charge voltage in terms of Tau and percentage of charge. Additionally, the same percentages also apply to the discharge time of a capacitor. For this example the capacitor was initially charged to 10V and will discharge 63.212% at 1τ leaving the capacitor voltage at 3.679V. Table **??** **??** has the compiled data and Figure 2.20 RC Differentiation, Generator and Capacitor Waveforms shows a graph of the generator and capacitor waveform voltages.

Once the capacitor voltages are calculated, the resistor voltage can be determined using Kirchhoff's Voltage Law. A tricky spot to pay attention to is the points where the generator voltage is vertical, 0 to 10v & 10v to 0. For example, you will need to do two Kirchhoff calculations at time equals 0.5mS, one where the generator voltage is equal to 0V and one where the generator voltage is equal to 10V. See Figure 2.21 RC Differentiation, Generator, Capacitor, and Resistor Waveforms.Notice how the resistor voltage goes negative, this is because at the instance that the generator voltage goes to zero, the capacitor voltage is still 10V, therefor to Kirchhoff, the resistor voltage must go negative.

$$V_R = V_{Gen} - V_C$$

Table 2.2: RC Circuits Capacitor Voltage

Capacitor Charge				
Tau	**Time**	**Generator Voltage**	**VC % of Charge**	**Capacitor Voltage**
0	**0μS**	**10V**	**0%**	**0V**
1	**50μS**	**10V**	**63.212%**	**6.321V**
2	**100μS**	**10V**	**86.466%**	**8.647V**
3	**150μS**	**10V**	**95.021%**	**9.502V**
4	**200μS**	**10V**	**98.168%**	**9.817V**
5	**250μS**	**10V**	**99.326%**	**9.933V**
$\geq 5\tau$	**500μS**	**10V**	**100%**	**10V**
250μS to 500μS the capacitor is fully charged to 10V				
Capacitor Discharge				
Tau	**Time**	**Generator Voltage**	**VC % of Discharge**	**Capacitor Voltage**
0	**500μS**	**0V**	**0%**	**10V**
1	**550μS**	**0V**	**63.212%**	**3.679V**
2	**600μS**	**0V**	**86.466%**	**1.353V**
3	**650μS**	**0V**	**95.021%**	**0.498V**
4	**700μS**	**0V**	**98.168%**	**0.183V**
5	**750μS**	**0V**	**99.326%**	**0.067V**
$\geq 5\tau$	**1mS**	**0V**	**100%**	**0V**

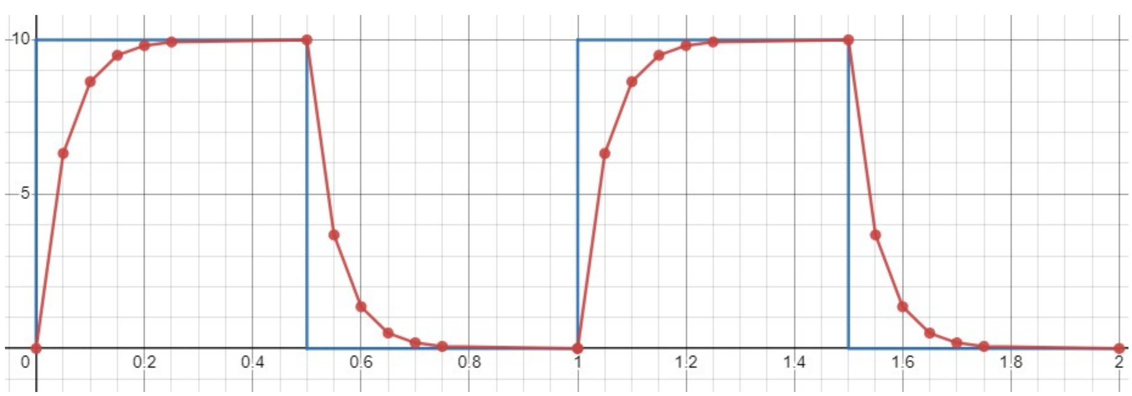


Figure 2.20: RC Differentiation, Generator and Capacitor Waveforms

Table 2.3: RC Circuits Capacitor & Resistor waveform voltages

Capacitor Charge					
Tau	**Time**	**Generator Voltage**	**VC % of Charge**	**Capacitor Voltage**	**Resistor Voltage**
0	0μS	10V	0%	0V	10V
1	50μS	10V	63.212%	6.321V	3.679V
2	100μS	10V	86.466%	8.647V	1.353V
3	150μS	10V	95.021%	9.502V	0.498V
4	200μS	10V	98.168%	9.817V	0.183V
5	250μS	10V	99.326%	9.933V	0.067V
$\geq 5\tau$	500μS	10V	100%	10V	0V
250μS to 500μS the capacitor is fully charged to 10V					
Capacitor Discharge					
Tau	**Time**	**Generator Voltage**	**VC % Discharge**	**Capacitor Voltage**	**Resistor Voltage**
0	500μS	0V	0%	10V	-10V
1	550μS	0V	63.212%	3.679V	-3.679V
2	600μS	0V	86.466%	1.353V	-1.353V
3	650μS	0V	95.021%	0.498V	-0.498V
4	700μS	0V	98.168%	0.183V	-0.183V
5	750μS	0V	99.326%	0.067V	-0.067V
$\geq 5\tau$	1mS	0V	100%	0V	0V

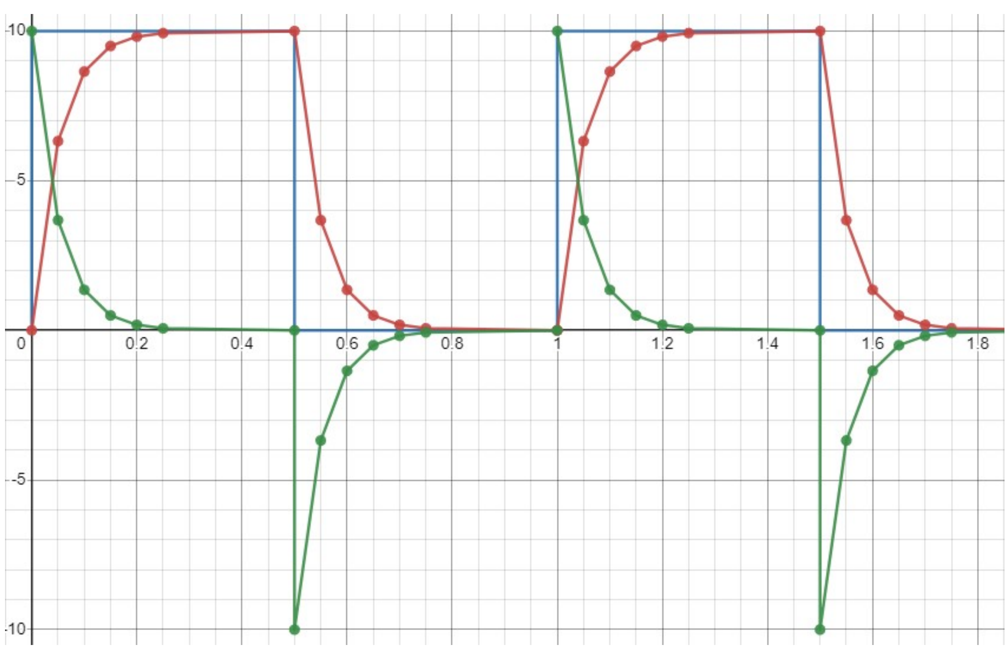


Figure 2.21: RC Differentiation, Generator, Capacitor, and Resistor Waveforms

2.12 Sine-waves and Instantaneous Voltage

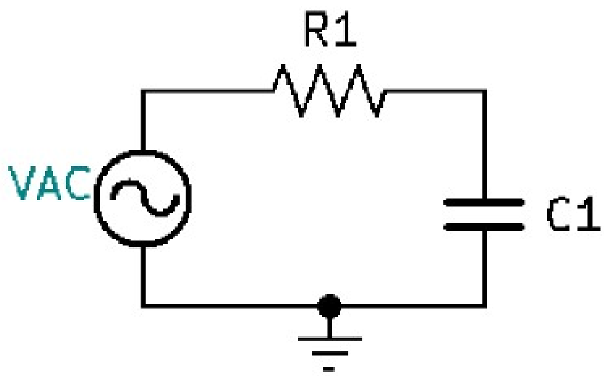


Figure 2.22: RC Circuit Sine-wave

2.12.1 Sine-wave analysis at resonance (F_C):

Consider Figure 2.22 RC Circuit Sine-wave.

Formulas:

- At F_C, $X_{C1} = R1$
- $X_{C1} = \frac{1}{2\pi F_C C1}$
- $R1 = \frac{1}{2\pi F_C C1}$
  - ✓ $F_C = \frac{1}{2\pi R1 C1}$
  - ✓ $Z_T = \sqrt{(R1)^2 + (X_{C1})^2}$, $\angle = \tan^{-1} \frac{X_{C1}}{R1}$
  - ✓ $I_T = \frac{V_{gen}}{Z_T}$
  - ✓ $V_{R1} = I_T \times R_1$
  - ✓ $V_{C1} = I_T \times X_{C1}$
  - ✓ $V_{inst} = V_{Max} \sin(360Ft \pm \theta)$

2.12.2 Sine-wave Instantaneous Example 1

Consider circuit Figure 2.22, and let's assume that V_{gen} is set to 10vp at 1Khz, and X_{C1} and $R1$ both are equal to $1K\Omega$.

Given:

- $V_{gen} = 10vp\ \angle 0°,\ 1Khz$
- $X_{C1} = 1K\Omega\ \angle -90°$
- $R1 = 1K\Omega\ \angle\ 0°$

Find: $Z_T,\ I_T,\ V_R,\ V_C,\ VGen_{inst},\ VR_{inst},\ VC_{inst}$

Solve:

$$Z_T = \sqrt{(R1)^2 + (X_{C1})^2}, \angle = \tan^{-1}\frac{X_{C1}}{R1}$$

$$Z_T = \sqrt{(1K)^2 + (1K)^2}, \angle = \tan^{-1}\frac{-1K}{1K}$$

$$\boxed{Z_T = 1.414K\Omega, \angle = -45°}$$

$$I_T = \frac{V_{gen}}{Z_T}$$

$$I_T = \frac{10vp\angle 0°}{1.414K\Omega\ \angle -45°}$$

$$\boxed{I_T = 7.071mAp\ \angle 45°}$$

$$V_{R1} = I_T \times R_1$$

$$V_{R1} = 7.071mAp\ \angle 45° \times 1K\Omega\ \angle\ 0°$$

$$\boxed{V_{R1} = 7.071vp\ \angle 45°}$$

$$V_{C1} = I_T \times X_C 1$$

$$V_{C1} = 7.071 mAp \angle 45^\circ \times 1K\Omega \angle -90^\circ$$

$$\boxed{V_{C1} = 7.071 vp \angle{-45^\circ}}$$

$$V_{inst} = V_{Max} \sin(360Ft \pm \theta)$$

$$\boxed{VGen_{inst} = 10vp \sin(360 \times 1Khz \times time \ \pm \ 0^\circ)}$$

$$\boxed{VR_{inst} = 7.071vp \sin(360 \times 1Khz \times time \ + \ 45^\circ)}$$

$$\boxed{VC_{inst} = 7.071vp \sin(360 \times 1Khz \times time \ - \ 45^\circ)}$$

Table 2.4: Instantaneous Voltages for Example 1. section 2.12.2

Time	VGen	VR1	VC1
0μS	0.000vp	5.000vp	-5.000vp
50μS	3.090vp	6.300vp	-3.210vp
100μS	5.878vp	6.984vp	-1.106vp
150μS	8.090vp	6.984vp	1.106vp
200μS	9.511vp	6.300vp	3.210vp
250μS	10.00vp	5.000vp	5.000vp
300μS	9.511vp	3.210vp	6.300vp
350μS	8.090vp	1.106vp	6.984vp
400μS	5.878vp	-1.106vp	6.984vp
450μS	3.090vp	-3.210vp	6.300vp
500μS	0.000vp	-5.000vp	5.000vp
550μS	-3.090vp	-6.300vp	3.210vp
600μS	-5.878vp	-6.984vp	1.106vp
650μS	-8.090vp	-6.984vp	-1.106vp
700μS	-9.511vp	-6.300vp	-3.210vp
750μS	-10.00vp	-5.000vp	-5.000vp
800μS	-9.511vp	-3.210vp	-6.300vp
850μS	-8.090vp	-1.106vp	-6.984vp
900μS	-5.878vp	1.106vp	-6.984vp
950μS	-3.090vp	3.210vp	-6.300vp
1mS	0.000vp	5.000vp	-5.000vp

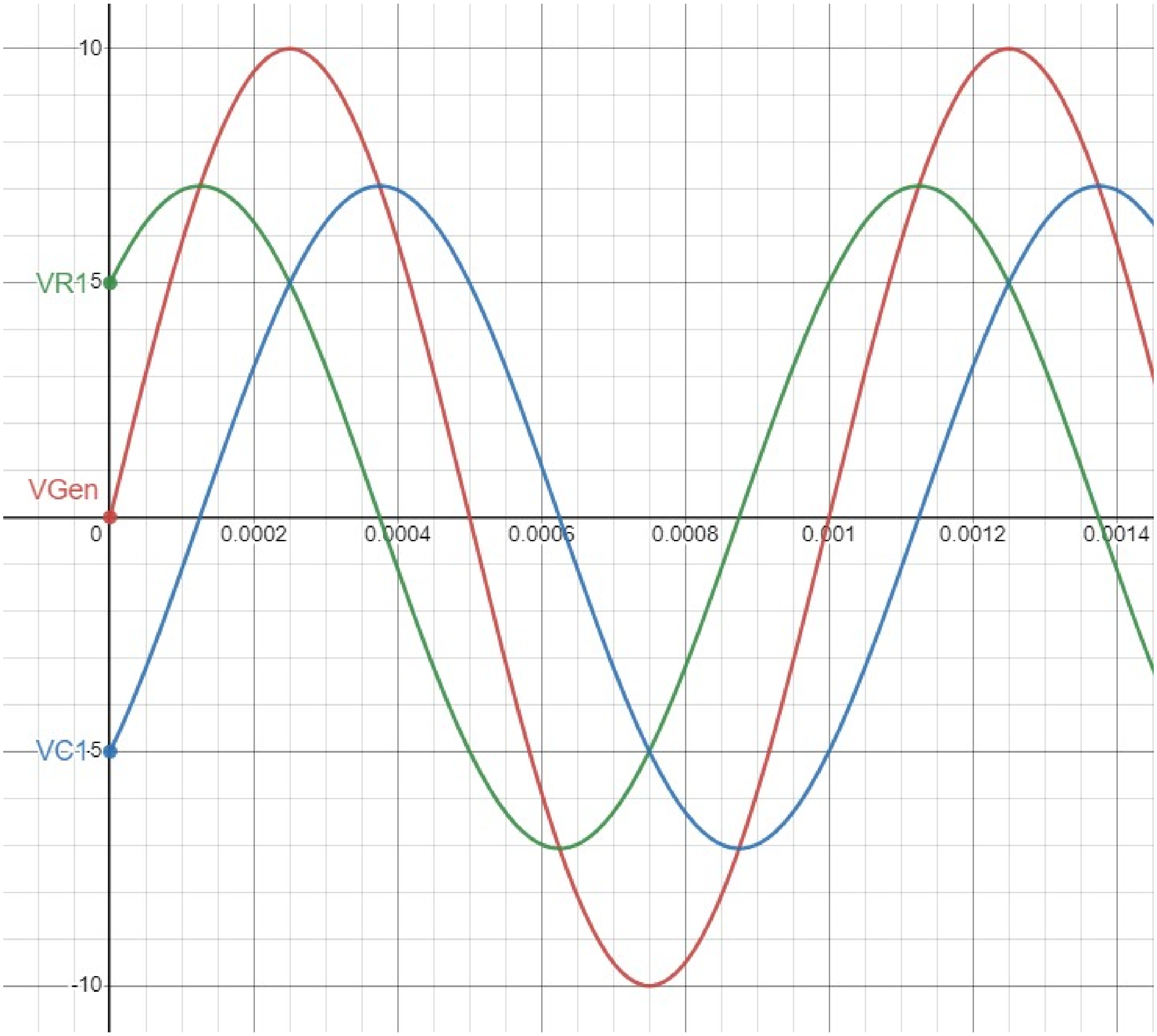


Figure 2.23: RC Circuit Instantaneous Waveforms Example 1. section 2.12.2

2.12.3 Sine-wave Instantaneous Example 2

What happens if the resistance of the previous example is doubled?

Given:

- $V_{gen} = 10vp \angle 0°,\ 1Khz$
- $X_{C1} = 1K\Omega \angle -90°$
- $R1 = 2K\Omega \angle 0°$

Find: $Z_T,\ I_T,\ V_R,\ V_C,\ VGen_{inst},\ VR_{inst},\ VC_{inst}$

Solve:

$$Z_T = \sqrt{(R1)^2 + (X_{C1})^2}, \angle = \tan^{-1}\frac{X_{C1}}{R1}$$

$$Z_T = \sqrt{(2K)^2 + (1K)^2}, \angle = \tan^{-1}\frac{-1K}{2K}$$

$$\boxed{Z_T = 2.236K\Omega, \angle = -26.565°}$$

$$I_T = \frac{V_{gen}}{Z_T}$$

$$I_T = \frac{10vp\angle 0°}{2.236K\Omega \angle -26.565°}$$

$$\boxed{I_T = 4.472mAp \angle 26.565°}$$

$$V_{R1} = I_T \times R_1$$

$$V_{R1} = 4.472mAp \angle 26.565° \times 2K\Omega \angle 0°$$

$$\boxed{V_{R1} = 8.944vp \angle 26.565°}$$

$$V_{C1} = I_T \times X_C 1$$

$$V_{C1} = 4.472mAp \angle 26.565^\circ \times 1K\Omega \angle -90^\circ$$

$$\boxed{V_{C1} = 4.472vp \angle{-63.435^\circ}}$$

$$V_{inst} = V_{Max} \sin(360Ft \pm \theta)$$

$$\boxed{VGen_{inst} = 10vp \sin(360 \times 1Khz \times time \ \pm \ 0^\circ)}$$

$$\boxed{VR_{inst} = 8.944vp \sin(360 \times 1Khz \times time \ + \ 26.565^\circ)}$$

$$\boxed{VC_{inst} = 4.472vp \sin(360 \times 1Khz \times time \ - \ 63.435^\circ)}$$

Table 2.5: Instantaneous Voltages for Example 2. section 2.12.3

Time	**VGen**	**VR1**	**VC1**
0μS	0.000vp	4.000vp	-4.000vp
50μS	3.090vp	6.276vp	-3.186vp
100μS	5.878vp	7.938vp	-2.060vp
150μS	8.090vp	8.823vp	-0.733vp
200μS	9.511vp	8.844vp	0.666vp
250μS	10.00vp	8.000vp	2.000vp
300μS	9.511vp	6.372vp	3.138vp
350μS	8.090vp	4.121vp	3.969vp
400μS	5.878vp	1.466vp	4.412vp
450μS	3.090vp	-1.332vp	4.422vp
500μS	0.000vp	-4.000vp	4.000vp
550μS	-3.090vp	-6.276vp	3.186vp
600μS	-5.878vp	-7.938vp	2.060vp
650μS	-8.090vp	-8.823vp	0.733vp
700μS	-9.511vp	-8.844vp	-0.666vp
750μS	-10.00vp	-8.000vp	-2.000vp
800μS	-9.511vp	-6.372vp	-3.138vp
850μS	-8.090vp	-4.121vp	-3.969vp
900μS	-5.878vp	-1.466vp	-4.412vp
950μS	-3.090vp	1.332vp	-4.422vp
1mS	0.000vp	4.000vp	-4.000vp

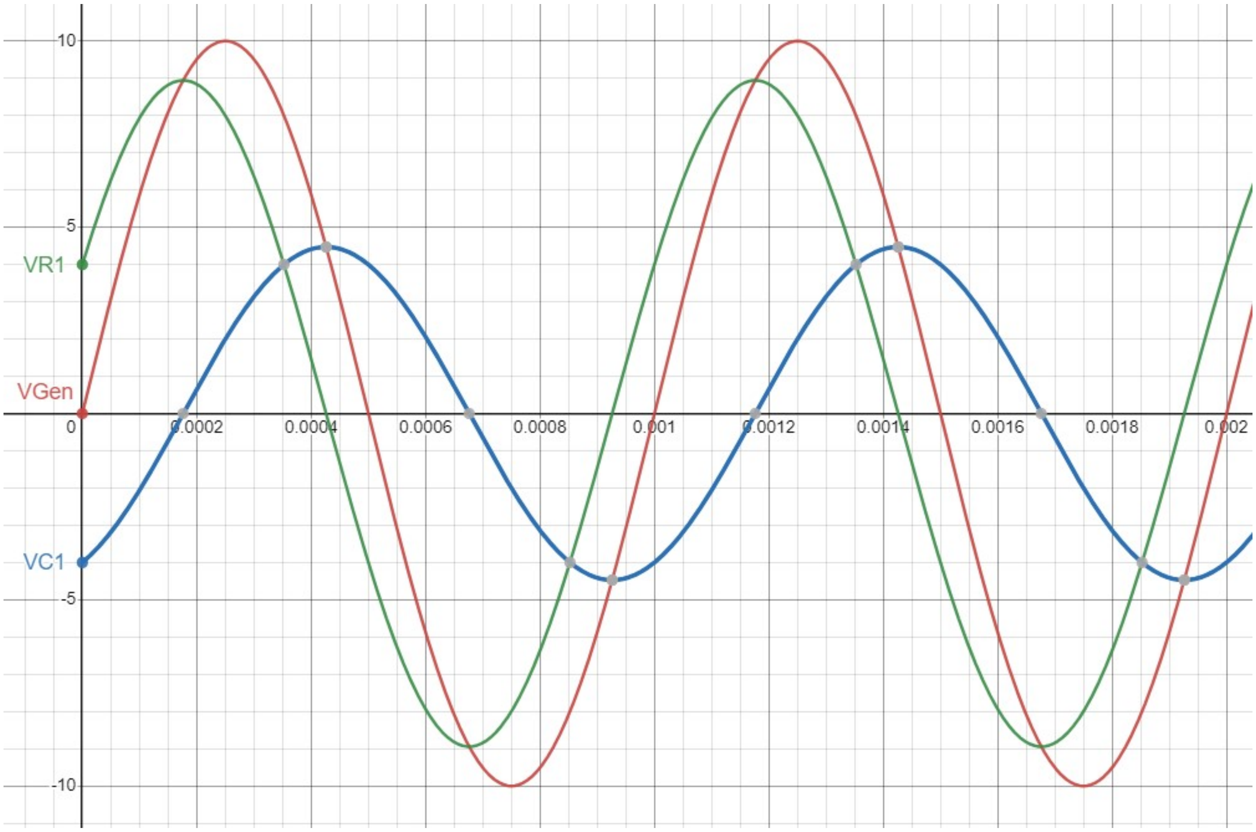


Figure 2.24: RC Circuit Instantaneous Waveforms Example 2. section 2.12.3

2.13 RC Circuits Questions

1.7 Graphs and Waveforms

1. On the graph, this axis represents the independent variable. (select all that apply)

   ☐ X-Axis

   ☐ Horizontal-Axis

   ☐ Y-Axis

   ☐ Vertical-Axis

2. This axis on the graph is used to represent the dependent variable. (select all that apply)

   ☐ X-Axis

   ☐ Horizontal-Axis

   ☐ Y-Axis

   ☐ Vertical-Axis

3. The point on the graph, where the two axis intersect is called the ____________ .

4. A ____________ waveform is a type of waveform that repeats its shape over regular intervals of time.

5. ____________ waveforms do not exhibit a regular and repeating pattern over time.

6. Common examples of periodic waveforms include which of the following waveforms?

   ☐ Sine

   ☐ Ramp

   ☐ Noise

   ☐ Sawtooth

   ☐ Exponetial

2.3 Pulse Waveform Characteristics

Matching:

1. ____ Period
2. ____ PW
3. ____ PS
4. ____ PRF
5. ____ DC%
6. ____ t_r
7. ____ t_f
8. ____ APA
9. ____ Tilt%
10. ____ AWV

(a) $=\frac{(+APA \times PW)+(-APA \times PS)}{Period}$

(b) The time it takes for one cycle.

(c) $=\frac{Vmax-Vmin}{APA} \times 100$

(d) Time High/On

(e) $=\frac{Vmax+Vmin}{2}$

(f) Time Low/Off

(g) The time required for the voltage to go from 90% to 10% of APA.

(h) $=\frac{1}{Period}$

(i) $=\frac{PW}{Period} \times 100$

(j) The time required for the voltage to go from 10% to 90% of APA.

Questions:

1. The ____________ ____________ waveform has perfectly vertical leading and lagging edges and perfectly flat tops and bottoms.
2. PW and PS are measured at ____________ of the waveform amplitude.
3. What is the DC% of a 1Khz waveform having with a pulse width equal to 250μS? Answer ____________ %.
4. What is the APA of a pulse that has a Vmax equal to 1.5V and a Vmin equal to 1V? Answer ____________ V.
5. What is the Tilt% of a pulse that has a Vmax equal to 1.5V and a Vmin equal to 1V? Answer ____________ %.
6. What is the AWV if +APA is equal to 10V and -APA is equal to -8V and PW is equal to 10μS and PS is equal to 20μS? Answer ____________ V.

1.8 Frequency Synthesis & Analysis

1. ____________ ____________ involves combining two or more signals to create a new waveform.
2. Select each true statement concerning Harmonics:
   - ☐ A harmonic is a multiple of the fundamental.
   - ☐ Harmonics are numbered according to their ratio to the fundamental.
   - ☐ The number of harmonics is infinite.
   - ☐ The 5^{th} harmonic frequency is equal to 5 times the Fundamental Frequency and the 5^{th} harmonic amplitude is equal to the Fundamentals Amplitude divided by 5.
3. A Perfect Square Wave is comprised of an infinite number of ____________ harmonic sine waves.
4. ____________ ____________ involves breaking down a complex waveform into its individual sinusoidal components or harmonics.
5. ____________ ____________ is a mathematical technique used to decompose a complex waveform into its individual sinusoidal components, representing different frequencies.
6. FFT stands for ____________ ____________ ____________ .
7. FFT can be used to analyze which of the following:
   - ☐ Analyze harmonics in power lines
   - ☐ Measure harmonic content and distortion in systems
   - ☐ Characterize noise in DC power supplies
   - ☐ Test impulse response of filters and systems
   - ☐ Analyze vibration

2.5 Waveform Distortion

1. The two types of square wave distortion that are useful for circuit frequency response analysis are ____________ and ____________ ____________ .

2.6 Tilt and Frequency Critical Low (FC_L)

1. ____________ capacitance will attenuate frequencies below FC_L.
2. ____________ represents the square wave distortion and attenuation of low frequencies.
3. FC_{Low} is equal to ____________ ____________ divided by 2π____________ .
4. What is FC_{Low} if the measured Tilt is 49.867% on a 630hz square wave.
   Answer: FC_{Low}=____________ hz.

2.7 Rise Time t_r and Frequency Critical High FC_H

1. The attenuation of high frequencies is caused by ____________ capacitance and will affect the ____________ ____________ of a measured square wave.
2. FC_{High} is equal to ____________ divided by ____________ .
3. If the measured Rise Time is 35μS, FC_{High} =____________ Khz.

2.8 Capacitor Charge Formula

1. At what percent of the Final Voltage will a capacitor charge in 1τ?
   Answer ____________ %
2. At what percent of the Final Voltage will a capacitor charge in 2τ?
   Answer ____________ %
3. At what percent of the Final Voltage will a capacitor charge in 3τ?
   Answer ____________ %
4. At what percent of the Final Voltage will a capacitor charge in 4τ?
   Answer ____________ %
5. At what percent of the Final Voltage will a capacitor charge in 5τ?
   Answer ____________ %

2.9 RC Circuits Stability, Vmax & Vmin Calculations

Matching:

1. ____ $V_{Capacitor}$
2. ____ $Stability_{Cycles}$
3. ____ V_{Max}
4. ____ V_{Min}

(a) $= V_{gen+} - V_{Max}$

(b) $= \frac{V_{gen+}}{1+e^{\frac{-PW}{RC}}}$

(c) $= \frac{5\tau}{Period}$

(d) $= V_{Final} - (V_{Final} - V_{Initial})e^{\frac{-t}{\tau}}$

Questions:

1. How many cycles will it take for the output to stabilize if the square wave frequency is 1Khz and τ is equal to 2mS? Answer ____________ cycles.
2. If the square wave voltage in question 1. is set at 0 to 10v, what is the stabilized V_{Max}? Answer V_{Max}=____________ V.
3. Considering the previous questions, What is V_{Min}? Answer V_{Min}= ____________ V.

2.10 RC Integration

1. Correctly select the **RC Integration Formula(s)** from the following:

   ☐ $RC \geq (10 \times PW)$
   ☐ $RC \leq (10 \times PW)$
   ☐ $PW \geq \frac{RC}{10}$
   ☐ $PW \leq \frac{RC}{10}$

2. Calculate the correct capacitance for an integrator circuit operating at 5Khz with a series resistance of 4.7KΩ. Select the correct answer:

   ☐ 0.220μF
   ☐ 0.210μF
   ☐ 0.120μF
   ☐ 0.110μF

3. For the previous question, Calculate the Cycles to Stabilization, V_{Max}, and V_{Min} with a generator voltage that is 0V to 10Vp.

   $Stabity =$ ____________ cycles
   $V_{Max} =$ ____________ V
   $V_{Min} =$ ____________ V

2.11 RC Differentiation

1. Correctly select the **RC Differentiation Formula(s)** from the following:

   ☐ $PW \geq (10 \times RC)$
   ☐ $PW \leq (10 \times RC)$
   ☐ $RC \geq \frac{PW}{10}$
   ☐ $RC \leq \frac{PW}{10}$

2. Calculate the correct capacitance for a differentiating circuit operating at 5Khz with a series resistance of 4.7KΩ. Select the correct answer:

   ☐ 0.022μF
   ☐ 0.021μF
   ☐ 2,200pF
   ☐ 2,100pF

2.12 Sine-waves and Instantaneous Voltage

Matching:

1. ____ At F_C,
2. ____ X_C
3. ____ F_C
4. ____ Z_T
5. ____ $\angle$
6. ____ I_T
7. ____ V_{Inst}

(a) $= V_{Max} sin(360Ft \pm \theta)$

(b) $\sqrt{(R)^2 + (X_C)^2}$

(c) $X_C = R$

(d) $= tan^{-1} \frac{X_C}{R}$

(e) $= \frac{V_T}{Z_T}$

(f) $= \frac{1}{2\pi RC}$

(g) $= \frac{1}{2\pi FC}$

Questions:

1. If a series RC circuit is operating at resonance with a generator voltage is $10vp$, $1Khz$ and circuit resistance equal to 2KΩ. Find the generator, resistor, and capacitor instantaneous voltage after 0.375mSec.

   $VGen_{0.375mS} =$ ____________ V
   $VR_{0.375mS} =$ ____________ V
   $VC_{0.375mS} =$ ____________ V

Week 3

Multi-Stage Amplifier: Design, Circuit Analysis, and Low-Frequency Response

3.1 Objectives:

Multi-Stage Amplifier Design and Analysis:

DC Biasing Calculations - Kirchhoff's and Thevenin Review.

- Kirchhoff's Laws: Students should be able to apply Kirchhoff's laws for DC biasing calculations in multi-stage amplifiers, considering voltage and current relationships in transistor amplifier circuits.
- Thevenin Equivalent: Understand and utilize Thevenin's theorem in transistor DC biasing calculations for simplifying complex circuits into simpler equivalent circuits.

AC Gain Calculations.

- AC Gain Analysis: Develop proficiency in calculating AC voltage gain for multi-stage amplifiers, considering the configuration and characteristics of each stage.

Load-Lines, Vout Max, and Vin Max.

- Load-Line Analysis: Understand load-line concepts and how to use them for analyzing the performance of multi-stage amplifiers.
- Vout Max Calculation: Calculate the maximum output voltage (Vout Max) for a multi-stage amplifier, considering various factors and load conditions.
- Vin Max Calculation: Calculate the maximum input voltage (Vin Max) that can be applied to a multi-stage amplifier without distortion or clipping.

Critical Frequencies.

- Low Critical Frequency (FCL): Understand and calculate the low critical frequency (FCL) for multi-stage amplifiers, considering the impact of coupling and bypass capacitors.
- High Critical Frequency (FCH): Understand and calculate the high critical frequency (FCH) for multi-stage amplifiers, taking into account the internal capacitances and parasitic elements in the amplifier stages.

By achieving these objectives, students will develop a comprehensive understanding of DC biasing calculations, AC gain, load-line analysis, maximum output and input voltage calculations, and critical frequencies for multi-stage amplifiers. These objectives aim to enhance their proficiency in designing and analyzing multi-stage amplifier circuits, particularly in terms of frequency response.

3.2 Biasing Configurations

Base Bias

3.3 Circuit Design, DC Biasing Stage 2:

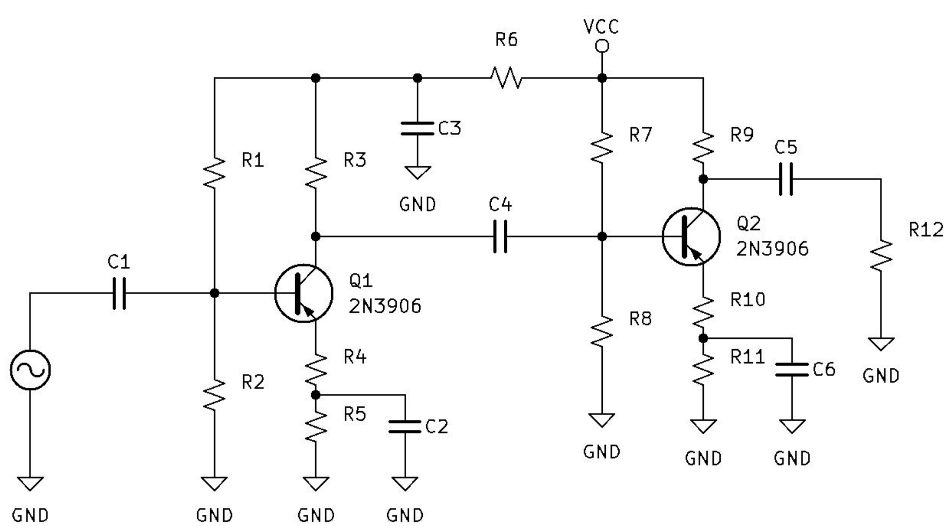


Figure 3.1: Two Stage Amplifier

Determine the Maximum Power for the 2N3906 Transistor

See Figure 3.2. P_D Total Device Dissipation.

Thermal Characteristics

Values are at T_A = 25°C unless otherwise noted.

Symbol	Parameter	Maximum			Unit
		2N3906(3)	MMBT3906(2)	PZT3906(3)	
P_D	Total Device Dissipation	625	350	1,000	mW
	Derate Above 25°C	5.0	2.8	8.0	mW/°C
$R_{\theta JC}$	Thermal Resistance, Junction to Case	83.3			°C/W
$R_{\theta JA}$	Thermal Resistance, Junction to Ambient	200	357	125	°C/W

Figure 3.2: Maximum Power for the 2n3906 transistor

DC Redraw for Stage 2:

Complete a DC redraw of the second stage of the amplifier making note of the DC biasing voltage polarities for each resistor and the transistor. See Figure 3.3.

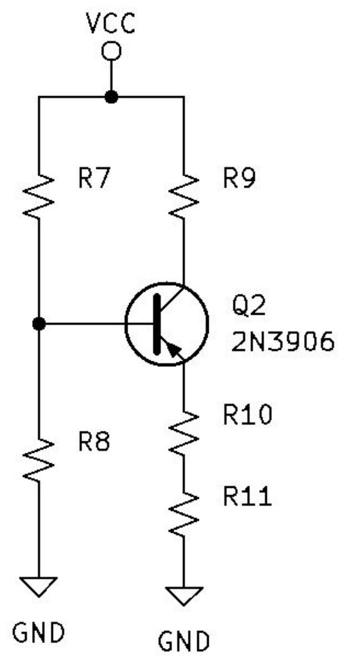


Figure 3.3: Second Stage DC Redraw

Determine Optimized Voltages and Transistor Power:

Optimize the biasing voltage using the 45, 45, 10 VCC ratios.

Voltage Optimization:

$V_{RC} = 45\% VCC$

$VCE = 45\%VCC$

$V_{RE} = 10\%VCC$

Make the transistor power ≈ half of the Maximum Power.

Q2 Power:

$P_{Q2} = \frac{P_{MAX}}{2} = 0.5P_{MAX}$

Collector Current:

$P_{Q2} = IC \times V_{CE}$

$IC = \frac{P_{Q2}}{V_{CE}}$

$IC = \frac{P_{MAX}}{2V_{CE}}$

$IC = \frac{P_{MAX}}{2\times 45\%VCC}$

$IC = \frac{P_{MAX}}{0.9VCC}$

Base Current:

$IB = \frac{IC}{Beta}$

$Beta_{2N3906} \approx 200$

$IB = \frac{P_{MAX}}{0.9VCC\times 200}$

$IB = \frac{P_{MAX}}{180VCC}$

Emitter Current:

$IE = IB \times (Beta + 1)$

$IE = \frac{P_{MAX}}{180VCC} \times (200 + 1)$

$IE = \frac{P_{MAX}\times 201}{180VCC}$

$IE = \frac{67P_{MAX}}{60VCC}$

R9:

$R9 = \frac{V_{R8}}{I_{R8}}$

$R9 = \frac{45\%VCC}{I_C}$

$R9 = \frac{45\%VCC}{\frac{P_{MAX}}{0.9VCC}}$

$R9 = \frac{45\%VCC \times 0.9VCC}{P_{MAX}}$

$R9 = \frac{0.405VCC^2}{P_{MAX}}$

Power R9:

$P_{R9} = IC \times V_{R8}$

$P_{R9} = IC \times 45\%VCC$

$P_{Q2} = IC \times 45\%VCC$

$P_{R9} = P_{Q2}$

$P_{R9} = \frac{P_{MAX}}{2} = 0.5P_{MAX}$

RE = R10 + R11:

$RE = \frac{V_{RE}}{I_{R9}}$

$RE = \frac{V_{RE}}{I_E}$

$RE = \frac{10\%VCC}{I_E}$

$RE = \frac{10\%VCC}{\frac{67P_{MAX}}{60VCC}}$

$RE = \frac{6VCC^2}{67P_{MAX}}$

Power RE ($P_{R10} + P_{R11}$)

$P_{RE} = IE \times V_{R9}$

$P_{RE} = \frac{67P_{MAX}}{60VCC} \times 10\%VCC$

$P_{RE} = \frac{6.7P_{MAX} \times VCC}{60VCC}$

$P_{RE} = \frac{6.7P_{MAX}}{60}$

R8:

$$R8 = \frac{V_{R8}}{I_{R8}}$$

$$V_{R8} = V_{RE} + V_{BE}$$

$$V_{R8} = 10\%VCC + 0.7V$$

$$I_{R8} = IB \times 10$$

$$I_{R8} = \frac{P_{MAX}}{180VCC} \times 10$$

$$I_{R8} = \frac{10P_{MAX}}{180VCC}$$

$$I_{R8} = \frac{P_{MAX}}{18VCC}$$

$$R8 = \frac{10\%VCC+0.7V}{\frac{P_{MAX}}{18VCC}}$$

$$R8 = \frac{18VCC(10\%VCC+0.7V)}{P_{MAX}}$$

$$R8 = \frac{1.8VCC^2+12.6VCC}{P_{MAX}}$$

Power R8:

$$P_{R8} = I_{R8} \times V_{R8}$$

$$P_{R8} = \frac{P_{MAX}}{18VCC} \times (10\%VCC + 0.7V)$$

$$P_{R8} = \frac{P_{MAX}(10\%VCC+0.7V)}{18VCC}$$

R7:

$$R7 = \frac{V_{R7}}{I_{R7}}$$

$$V_{R7} = VCC - V_{R8}$$

$$V_{R7} = VCC - (10\%VCC + 0.7V)$$

$$V_{R7} = VCC - 10\%VCC - 0.7V$$

$$V_{R7} = 0.9VCC - 0.7V$$

$$I_{R7} = I_{R7} + IB$$

$$I_{R7} = \frac{P_{MAX}}{180VCC} + \frac{P_{MAX}}{18VCC}$$

$$I_{R7} = \frac{P_{MAX}}{180VCC} + \frac{10P_{MAX}}{180VCC}$$

$$I_{R7} = \frac{P_{MAX}+10P_{MAX}}{180VCC}$$

$$I_{R7} = \frac{P_{MAX}(1+10)}{180VCC}$$

$$I_{R7} = \frac{11P_{MAX}}{180VCC}$$

$$R7 = \frac{0.9VCC - 0.7V}{\frac{11P_{MAX}}{180VCC}}$$

$$R7 = \frac{180VCC(0.9VCC - 0.7V)}{11P_{MAX}}$$

$$R7 = \frac{162VCC^2 - 126VCC}{11P_{MAX}}$$

Power R7:

$$P_{R7} = I_{R7} \times V_{R7}$$

$$P_{R7} = \frac{11P_{MAX}}{180VCC} \times (0.9VCC - 0.7V)$$

$$P_{R7} = \frac{P_{MAX}(9.9VCC + 7.7V)}{180VCC}$$

3.4 Gain Calculations for Stage 2

R12

Make R12 ten times larger than R9.

$$R12 = R9 \times 10$$

Determine the Voltage Gain for Stage 2

$$\Delta V = \frac{V_{OUT}}{V_{IN}}$$

$$\Delta V = \frac{IC(R9//R12)}{IB(R10 + r'e)(B+1)}$$

$$\Delta V = \frac{IC}{IB} \times \frac{(R9//R12)}{(R10 + r'e)(B+1)}$$

$$\Delta V = \beta \times \frac{(R9//R12)}{(R10 + r'e)(B+1)}$$

$$\Delta V = \frac{\beta}{\beta+1} \times \frac{(R9//R12)}{(R10 + r'e)}$$

$$\Delta V = \alpha \times \frac{(R9//R12)}{(R10 + r'e)}$$

$$\Delta V = \frac{\alpha(R9//R12)}{(R10 + r'e)}$$

Calculating R10 to set the Desired Gain for Stage 2

Choose the desired gain, $\Delta V_{desired}$ and solve for a $R10$ value.

$$\Delta V_{desired} = \frac{\alpha(R9//R12)}{(R10+r'e)}$$

$$R10 + r'e = \frac{\alpha(R9//R12)}{\Delta V_{desired}}$$

$$R10 = \frac{\alpha(R9//R12)}{\Delta V_{desired}} - r'e$$

R11

$$RE = R10 + R11$$

$$R11 = RE - R10$$

3.5 Circuit Design, DC Biasing Stage 1:

DC Redraw for Stage 1:

Complete a DC redraw of the second stage of the amplifier making note of the DC biasing voltage polarities for each resistor and the transistor. See Figure 3.4.

Decoupling Resistor R6

The primary objective of the Decoupling Resistor R6 is to prevent oscillations by eliminating any positive feedback from the output of Q2 back to the input of Q1. C2 of Figure 3.1 will act like an AC short to ground allowing any potential feedback signal to be dropped on R6.

Additionally, R6 can be used to significantly lower the power for Q1 if needed.

Find the equivalent load resistance RL_{EqQ1} for Q1:

$$RL_{EqQ1} = ((R10 + r'e_{Q2}) \times (\beta_{Q2} + 1))//R7//R8$$

R3:

Make $R3$ ten times smaller than RL_{EqQ1}.

$$R3 = \frac{RL_{EqQ1}}{10}$$

Make P_{R3} equal to half the maximum power of Q1.

$$P_{R3} = \frac{PMAX_{Q1}}{2}$$

Calculate I_{R3}.

$$P = I^2 \times R$$

$$P_{R3} = I_{R3}^2 \times R3$$

$$I_{R3}^2 = \frac{P_{R3}}{R3}$$

$$I_{R3} = \sqrt{\frac{P_{R3}}{R3}}$$

Calculate V_{R3}.

$$P = \frac{V^2}{R}$$

$$P_{R3} = \frac{V_{R3}^2}{R3}$$

$$V_{R3}^2 = P_{R3} \times R3$$

$$V_{R3} = \sqrt{P_{R3} \times R3}$$

Q1:

$$IC = I_{R3}$$

$$IB = \frac{IC}{\beta}$$

$$IE = IB \times (\beta + 1)$$

Make V_{CE} of Q1 equal to V_{R3}.

$$V_{CE} = V_{R3}$$

$$P_{Q1} = P_{R3} = \frac{PMAX_{Q1}}{2}$$

RE:

$$RE = R4 + R5$$

Following the 45, 45, 10 optimization, Find VRE.

$$V_{CE} = .45X$$

$$X = \frac{V_{CE}}{.45}$$

$$V_{RE} = .10X$$

$$X = \frac{V_{RE}}{.10}$$

if X=X, then:

$$\frac{V_{CE}}{.45} = \frac{V_{RE}}{.10}$$

$$V_{RE} = \frac{.10V_{CE}}{.45}$$

$$V_{RE} = \frac{V_{CE}}{4.5}$$

R2:

$$V_{R2} = V_{BE} + V_{RE}$$

$$V_{R2} = 0.7v + V_{RE}$$

Make IR2 ten times larger than the base current of Q1.

$$I_{R2} = IB \times 10$$

$$R2 = \frac{V_{R2}}{I_{R2}}$$

R1:

Find VR1.

$$V_{R1} + V_{R2} = V_{R3} + V_{CE} + V_{RE}$$

$$V_{R1} = V_{R3} + V_{CE} + V_{RE} - V_{R2}$$

Calculate IR1.

$$I_{R1} = I_{R2} + I_B$$

Solve for R1.

$$R1 = \frac{V_{R1}}{I_{R1}}$$

R6:

Find VR6.

$$V_{R2} + V_{R1} + V_{R6} - V_{CC} = 0$$

$$V_{R6} = V_{CC} - V_{R2} - V_{R1}$$

Calculate IR6.

$$I_{R6} = I_{R1} + I_{R3}$$

Solve for R6.

$$R6 = \frac{V_{R6}}{I_{R6}}$$

3.6 Gain Calculations for Stage 1

Find the equivalent load resistance RL_{EqQ1} for Q1:

$$RL_{EqQ1} = ((R10 + r'e_{Q2}) \times (\beta_{Q2} + 1))//R7//R8$$

Find r'e for Q1

$$r'e = \frac{0.026}{IE}$$

Determine the Voltage Gain for Stage 1

$$\Delta V = \frac{V_{OUT}}{V_{IN}}$$

$$\Delta V = \frac{IC(R3//RL_{EqQ1})}{IB(R4{+}r'e)(B{+}1)}$$

$$\Delta V = \frac{IC}{IB} \times \frac{(R3//RL_{EqQ1})}{(R4{+}r'e)(B{+}1)}$$

$$\Delta V = \beta \times \frac{(R3//RL_{EqQ1})}{(R4{+}r'e)(B{+}1)}$$

$$\Delta V = \frac{\beta}{\beta+1} \times \frac{(R3//RL_{EqQ1})}{(R4{+}r'e)}$$

$$\Delta V = \alpha \times \frac{(R3//RL_{EqQ1})}{(R4{+}r'e)}$$

$$\Delta V = \frac{\alpha(R3//RL_{EqQ1})}{(R4{+}r'e)}$$

Calculating R4 to set the Desired Gain for Stage 1

Choose the desired gain, $\Delta V_{desired}$ and solve for a $R4$ value.

$$\Delta V_{desired} = \frac{\alpha(R3//RL_{EqQ1})}{(R4{+}r'e)}$$

$$R4 + r'e = \frac{\alpha(R3//RL_{EqQ1})}{\Delta V_{desired}}$$

$$R4 = \frac{\alpha(R3//RL_{EqQ1})}{\Delta V_{desired}} - r'e$$

R5

$$RE = R4 + R4$$

$$R5 = RE - R4$$

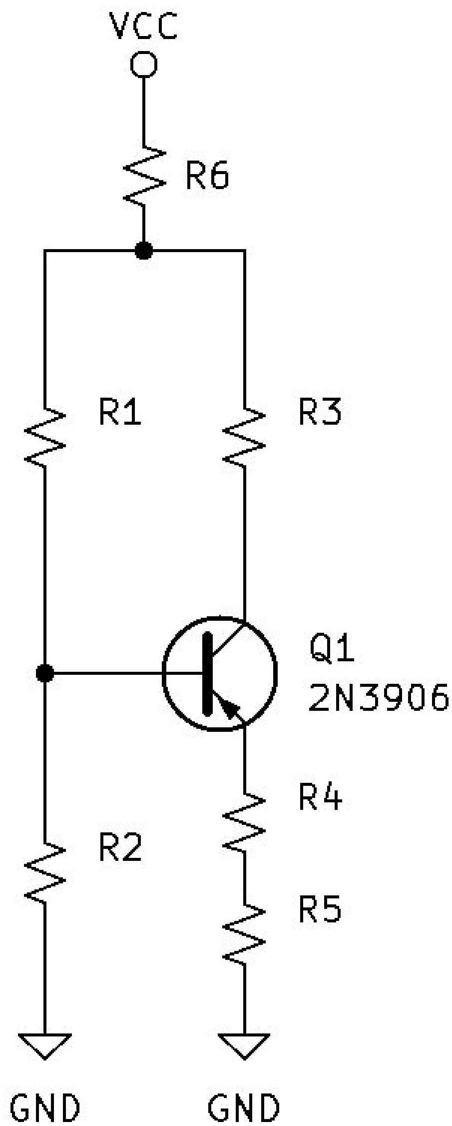


Figure 3.4: First Stage DC Redraw with polarities

Q2, DC Loadline

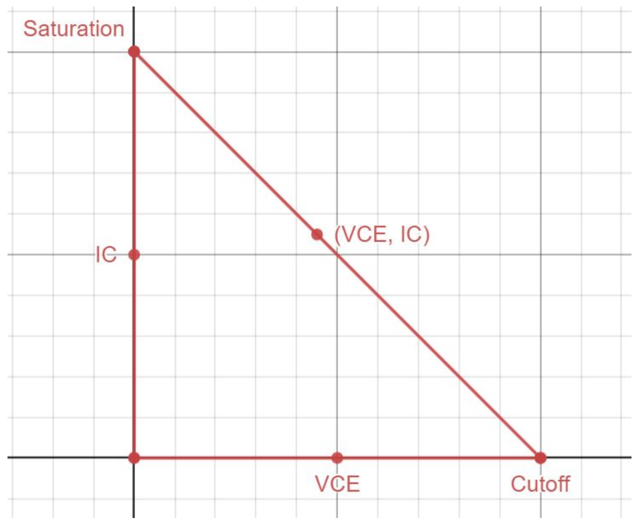


Figure 3.5: Second Stage DC Loadline

IC Saturation in terms of VCC and Max Power:

$$IC_{Sat} = \frac{VCC}{R9+RE}$$

$$IC_{Sat} = \frac{VCC}{\frac{0.405VCC^2}{P_{MAX}} + \frac{6VCC^2}{67P_{MAX}}}$$

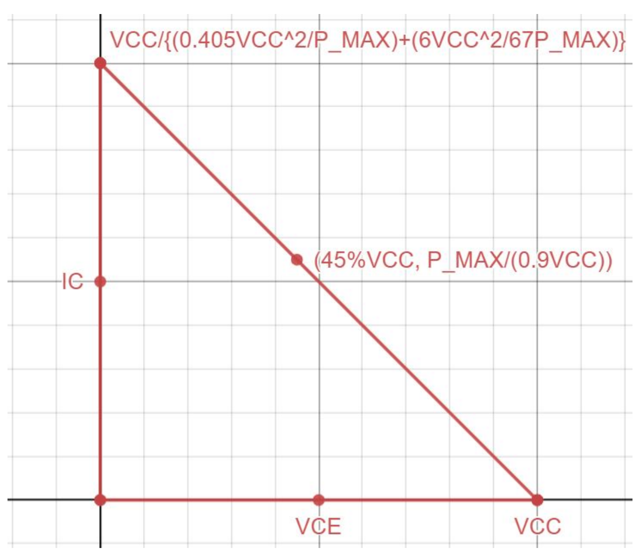


Figure 3.6: Second Stage DC Loadline in Terms of VCC and Max Power

Q2, AC Loadline

Because the load resistor R12 is ten times larger than R9 and the biasing is optimized, $Vout_{MAXp}$ will be just less than VCE.

$$Vout_{MAXp} \approx 0.9(V_{CE\_Q2})$$

$$V_{CE\_AC\_Cutoff} = V_{CE\_Q2} + I_{R9}(R9//R12) \qquad IC_{AC\_Sat} = I_{R9} + \frac{V_{CE\_Q2}}{R9//R12}$$

$$Vout_{MAXp} = V_{CE\_AC\_Cutoff} - V_{CE\_Q2}$$

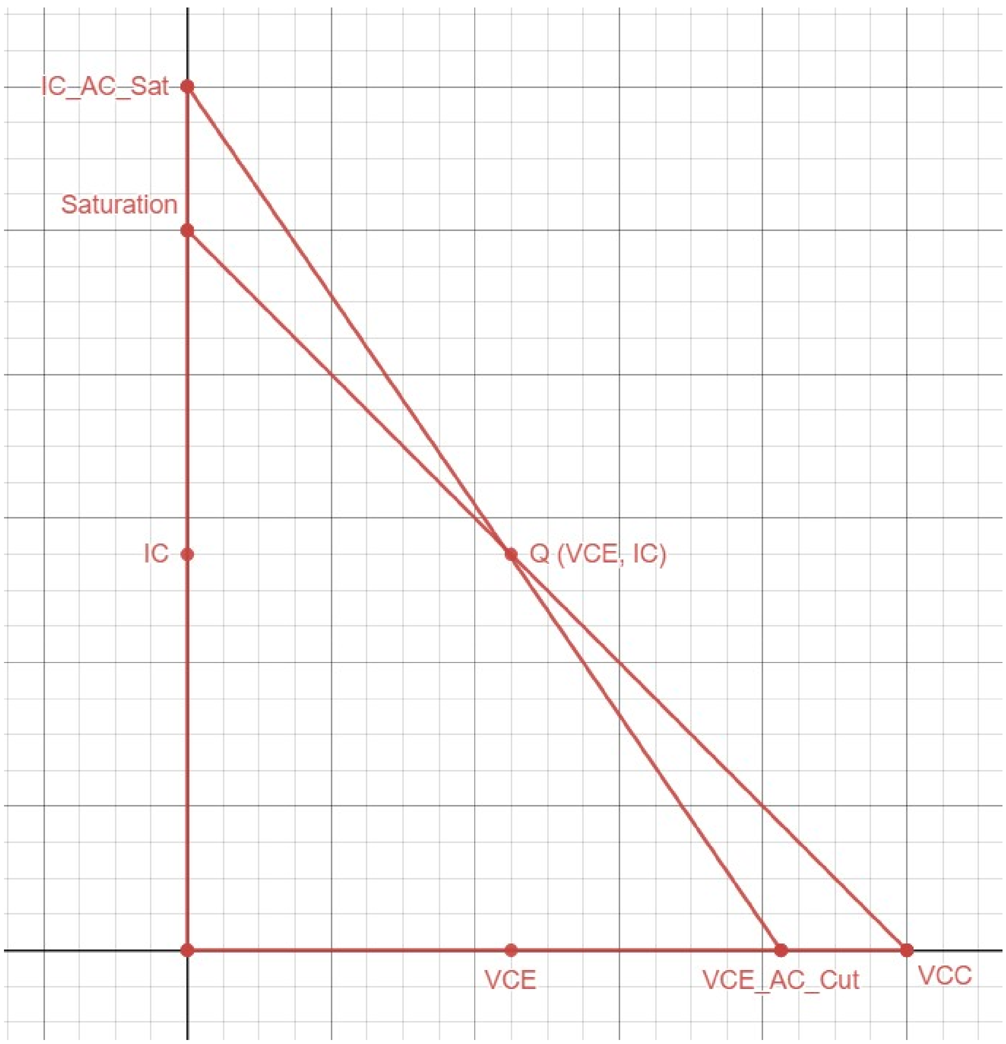


Figure 3.7: Second Stage DC & AC Loadline

Q1, Loadline

Because of the Decoupling Resistor R6, the DC Load Line is not Optimized at 45, 45 10 and will need to be calculated. However, because the equivalent load resistance of Q1 is ten times larger than R3, stage 1 maximum peak output should also be just below Q1's VCE.

$$Vout_{MaxP\_Q1} \approx 0.9(V_{CE\_Q1})$$

Additionally, if the voltage gain of the first stage is approximately equal to or less than the voltage gain of the second stage, the actual maximum peak voltage out will be determined by the second stage. This can be verified by using the following steps:

1. Divide the second stage Vout max peak voltage by the voltage gain of the second stage.

   $$X = \frac{Vout_{Q2\_MAXp}}{\Delta V_{Q2}}$$

2. Verify that Q1 VCE is greater than X.

   ✓ $V_{CE\_Q1} > X$

3. If X is greater than or equal to Q1 VCE, biasing could be adjusted by redesigning the circuit to raise Q1's VCE and lower the decoupling voltage. However, it may be quicker and easier to lower Q1's voltage gain and raise Q2's voltage gain by the same factor.

3.7 Circuit Analysis

Universal Biasing

The Universal Bias Circuit can be analyzed using either Thevenin or Kirchhoff analysis. Kirchhoff's analysis is based on Kirchhoff's Voltage and Current Laws and will always produce an accurate analysis. Thevenin Analysis is based on Thevenin's Theorem and produces accurate analysis as long as certain circuit parameters are met.

Universal Bias, Thevenin Analysis

Thevenin Analysis converts a complex circuit into a simple, easier-to-analyze, series circuit see Figure 3.8.

Thevenin Analysis Steps:

Find the Thevenin Voltage V_{Thev}.

V_{Thev} is the maximum voltage that the load will see. Consider Q2 and RE as the load, see Figure 3.1.

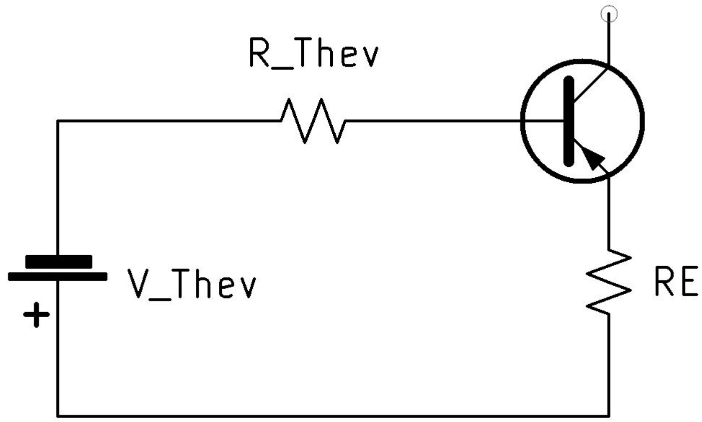


Figure 3.8: Thevenin Equivalent Circuit for Q2

Imagine that Q2 and RE (R10 and R11) are removed from the circuit.

R7 and R8 now make a series circuit, and V_{Thev} is equal to V_{R8}. See Figure 3.9

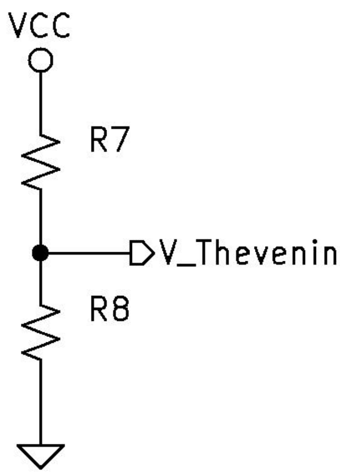


Figure 3.9: Thevenin Voltage

Find the Thevenin Resistance R_{Thev}.

R_{Thev} is equal to the total resistance that the load sees, for our example, R7 in parallel with R8.

$R_{Thev} = R7//R8$

Find the base current IB of the transistor.

Write a loop equation:

$V_{Thev} - V_{R\_Thev} - V_{BE} - V_{RE} = 0$

$$V_{R\_Thev} + V_{RE} = V_{Thev} - V_{BE}$$
$$IB(R\_Thev) + IE(RE) = V_{Thev} - V_{BE}$$
$$IB(R\_Thev) + (IB(\beta+1))(RE) = V_{Thev} - V_{BE}$$
$$IB\left((R\_Thev) + (\beta+1)(RE)\right) = V_{Thev} - V_{BE}$$

$$IB = \frac{V_{Thev}-V_{BE}}{(R\_Thev)+(\beta+1)(RE)}$$

Verify that IR8 is greater than or equal to IB. If IR8 is not greater than or equal to ten times IB, the circuit will need to be analyzed using Kirchhoff Analysis.

✓ $I_{R8} \geq 10 \times IB$

Universal Bias, Kirchhoff Analysis

Kirchhoff's Analysis of Universal Biasing requires two loop equations.

Kirchhoff Loop Equation 1 (solve for IR8 in terms of IR8 and IB):

$$VCC - V_{R7} - V_{R8} = 0$$
$$VCC - I_{R7}(R7) - I_{R8}(R8) = 0$$
$$VCC - (IB + I_{R8})(R7) - I_{R8}(R8) = 0$$
$$VCC - IB(R7) - I_{R8}(R7) - I_{R8}(R8) = 0$$
$$VCC - IB(R7) - I_{R8}(R7 + R8) = 0$$
$$I_{R8}(R7 + R8) = VCC - IB(R7)$$

✓ $I_{R8} = \frac{VCC-IB(R7)}{(R7+R8)}$

Kirchhoff Loop Equation 2 (solve for IR8 in terms of IB):

$$VCC - V_{R7} - V_{BE} - V_{RE} = 0$$
$$VCC - I_{R7}(R7) - 0.7 - IE(RE) = 0$$
$$VCC - (IB + I_{R8})(R7) - 0.7 - IB(\beta+1)(RE) = 0$$
$$VCC - IB(R7) - I_{R8}(R7) - 0.7 - IB(\beta+1)(RE) = 0$$
$$VCC - IB(R7) - 0.7 - IB(\beta+1)(RE) = I_{R8}(R7)$$
$$I_{R8}(R7) = VCC - IB(R7) - 0.7 - IB(\beta+1)(RE)$$
$$I_{R8}(R7) = VCC - 0.7 - IB(R7 + (\beta+1)(RE))$$

✓ $I_{R8} = \frac{VCC-0.7-IB(R7+(\beta+1)(RE))}{R7}$

Use the substitution method to solve for IB by setting each equation equal to the other.

$$\frac{VCC-IB(R7)}{(R7+R8)} = \frac{VCC-0.7-IB(R7+(\beta+1)(RE))}{R7}$$

$$R7(VCC - IBR7) = (R7 + R8)(VCC - 0.7 - IB(R7 + (\beta + 1)(RE))$$

$$R7VCC - IBR7^2 = (R7 + R8)(VCC - 0.7 - IBR7 - IB(\beta + 1)(RE))$$

$$R7VCC - IBR7^2 = R7VCC - R7(0.7) - IBR7^2 - IBR7(\beta+1)(RE) + R8VCC - R8(0.7) - IBR7R8 - IBR8(\beta + 1)(RE)$$

$$0 = -R7(0.7) - IBR7(\beta + 1)(RE) + R8VCC - R8(0.7) - IBR7R8 - IBR8(\beta + 1)(RE)$$

$$IBR7(\beta+1)(RE) + IBR7R8 + IBR8(\beta+1)(RE) = -R7(0.7) + R8VCC - R8(0.7)$$

$$IB\{R7(\beta + 1)(RE) + R7R8 + R8(\beta + 1)(RE)\} = R8(VCC - 0.7) - R7(0.7)$$

$$IB = \frac{R8(VCC-0.7)-R7(0.7)}{(R7)(\beta+1)(RE)+R7R8+R8(\beta+1)(RE)}$$

✓ $$IB = \frac{R8(VCC-0.7)-R7(0.7)}{(R7)(\beta+1)(RE)+R8(R7+(\beta+1)(RE))}$$

Universal Bias with Decoupling Resistor, Kirchhoff Analysis

Kirchhoff's Analysis of Universal Biasing requires two loop equations.

Kirchhoff Loop Equation 1 (solve for IR2 in terms of IB):

$$VCC - VR6 - VR1 - VR2 = 0$$
$$VCC - I_{R6}R6 - I_{R1}R1 - I_{R2}R2 = 0$$
$$VCC - (IC + I_{R1})R6 - I_{R1}R1 - I_{R2}R2 = 0$$
$$VCC - (IB(\beta) + I_{R2} + IB)R6 - (I_{R2} + IB)R1 - I_{R2}R2 = 0$$
$$VCC - IB(\beta)R6 - I_{R2}R6 - IBR6 - I_{R2}R1 - IBR1 - I_{R2}R2 = 0$$
$$I_{R2}R6 + I_{R2}R1 + I_{R2}R2 = VCC - IB(\beta)R6 - IBR6 - IBR1$$
$$I_{R2}(R6 + R1 + R2) = VCC - (IB)((\beta)R6 + R6 + R1)$$

✓ $I_{R2} = \frac{VCC-(IB)((\beta)R6+R6+R1)}{R6+R1+R2}$

Kirchhoff Loop Equation 2 (solve for IR2 in terms of IB):

$$-V_{BE} - V_{RE} + V_{R2} = 0$$
$$-0.7V - (IE)RE + I_{R2}R2 = 0$$
$$-0.7V - IB(\beta + 1)RE + I_{R2}R2 = 0$$
$$I_{R2}R2 = 0.7V + IB(\beta + 1)RE$$

✓ $I_{R2} = \frac{0.7V+IB(\beta+1)RE}{R2}$

Use the substitution method to solve for IB by setting each equation equal to the other.

$$\frac{VCC-(IB)((\beta)(R6+R6+R1)}{R6+R1+R2} = \frac{0.7V+IB(\beta+1)RE}{R2}$$

$$R2[VCC - (IB)((\beta)(R6 + R6 + R1)] = (R6 + R1 + R2)(0.7V + IB(\beta + 1)(RE)$$

$$VCC(R2) - (IB)(R2)((\beta)R6 + R6 + R1) = 0.7V(R6 + R1 + R2) + IB(\beta + 1)(RE)(R6 + R1 + R2)$$

$$VCC(R2) - 0.7V(R6 + R1 + R2) = (IB)(R2)((\beta)R6 + R6 + R1) + IB(\beta + 1)(RE)(R6 + R1 + R2)$$

$$VCC(R2) - 0.7V(R6+R1+R2) = (IB)\{(R2)((\beta)R6+R6+R1)+(\beta+1)(RE)(R6+R1 + R2)\}$$

$$\frac{VCC(R2)-0.7V(R6+R1+R2)}{\{(R2)((\beta)R6+R6+R1)+(\beta+1)(RE)(R6+R1+R2)\}} = (IB)$$

✓ $IB = \frac{VCC(R2)-0.7V(R1+R2+R6)}{R2(R1+R6+(\beta\times R6))+RE(\beta+1)(R1+R2+R6)}$

➥ Find all resistor voltages and powers.

Q2, Power and Load-line Analysis

Find V_{CE_Q2}

$$V_{CE\_Q2} = VCC - V_{R9} - V_{R10} - V_{R11}$$

Find IC

$$IC_{R9} = \frac{V_{R9}}{R9}$$

Find P_{Q2}

$$P_{Q2} = V_{CE\_Q2} \times IC_{R9}$$

Find IC_{Sat_Q2}

$$IC_{Sat\_Q2} = \frac{VCC}{R9+R10+R11}$$

Find V_{cut_Q2}

$$V_{cut\_Q2} = VCC$$

Find $v_{cut\_ac\_Q2}$

$$v_{cut\_ac\_Q2} = V_{CE\_Q2} + IC_{R9}(R12//R9)$$

Find ic_{ac_Q2}

$$ic_{ac\_Q2} = IC_{Sat\_Q2} + \frac{V_{CE\_Q2}}{(R12//R9)}$$

Depending on the biasing (Q point), the smaller of the two following equations will represent $Vout_{MaxP\_Q2}$. If the amplifier is designed for a Z_{out} (R9) that is ten times smaller than the load (R12) and the biasing is optimized, Vout Max Peak will be approximately equal to 90% of VCE.

$$Vout_{MaxP\_Q2} \approx 0.9(V_{CE\_Q2})$$

Vout max peak actual will be the smaller of the following:

$$VoutP_{Max\_Q2} = v_{cut\_ac\_Q2} - V_{CE\_Q2}$$

OR

$$VoutP_{Max\_Q2} = V_{CE\_Q2}$$

Q1, Power and Load-line Analysis

Find V_{CE_Q1}, $V_{CE\_Q1} = VCC - V_{R6} - V_{R3} - V_{R4} - V_{R5}$

Find IC, $IC_{R3} = \frac{V_{R3}}{R3}$

Find P_{Q1}, $P_{Q1} = V_{CE\_Q1} \times IC_{R3}$

Find $VoutP_{Max\_Q1}$, $VoutP_{Max\_Q1} \approx 0.9(V_{CE\_Q1})$

Q2, Voltage Gain Analysis

Find ΔV_{Q2}, $\Delta V_{Q2} \approx \frac{R9}{R10}$

$$\Delta V_{Q2} = \frac{V_{OUT}}{V_{IN}}$$
$$\Delta V_{Q2} = \frac{IC(R9//R12)}{IB(\beta+1)(R10+r'e)}$$
$$\Delta V_{Q2} = \frac{IC(R9//R12)}{IE(R10+r'e)}$$
$$\Delta V_{Q2} = \alpha\frac{(R9//R12)}{R10+r'e}$$

Q1, Voltage Gain Analysis

Find ΔV_{Q1}, $\Delta V_{Q1} \approx \frac{R3}{R4}$

$$\Delta V_{Q1} = \frac{V_{OUT}}{V_{IN}}$$
$$\Delta V_{Q1} = \frac{IC(R3//R7//R8//((R10+r'e)(\beta+1))}{IB(\beta+1)(R4+r'e)}$$
$$\Delta V_{Q1} = \frac{IC(R3//R7//R8//((R10+r'e)(\beta+1))}{IE(R4+r'e)}$$
$$\Delta V_{Q1} = \alpha\frac{R3//R7//R8//((R10+r'e)(\beta+1))}{R4+r'e}$$

Total Voltage Gain Analysis

Find ΔV_{Total},

$$\Delta V_{Total} = \Delta V_{Q1} \times \Delta V_{Q2}$$

Maximum Peak Input Voltage

Find $VinP_{Max}$,

$$VinP_{Max} = \frac{VoutP_{Max}}{\Delta V_{Total}}$$

3.8 Frequency Response Low

Rules and Steps for Calculating Frequency Response Low

Rules:

1. Treat all capacitors like opens.
2. Find $R_{Thevenin}$ for each capacitor.
3. Calculate Fc_{Low_C} for each capacitor:

$$Fc_{Low\_C} = \frac{1}{2\pi Rthev_C C}$$

4. Calculate Fc_{LOW} Total

$$FC_{Low} = \sqrt{(FC_{C1})^2 + (FC_{C2})^2 + (FC_{C3})^2 ...}$$

Steps:

Calculate the Capacitor Thevenin Resistances for Figure 3.1.

$Rthev_{C1} = \{(R5 + R4 + r'e_{Q1})(\beta + 1)//R2//(R1 + R6)\} + R_{Gen}$

$Rthev_{C2} = \{(\frac{R2//(R1+R6)}{\beta+1}) + r'e_{Q1} + R4\}//R5$

$Rthev_{C3} = R6//\{R1 + (R2//((\beta + 1)(r'e_{Q1} + R4 + R5))\}$

$Rthev_{C4} = \{((R11 + R10 + r'e_{Q2})(\beta + 1))//R7//R8\} + \{(((((R5 + R4 + r'e_{Q1})(\beta + 1))//R2) + R1)//R6) + R3\}$

$Rthev_{C5} = R9 + R12$

$Rthev_{C6} = ((\frac{R7//R8}{\beta+1}) + r'e_{Q2} + R10)//R11$

Design: Calculating capacitor values for a desired Frequency Critical Low

Choose a desired Frequency Critical Low $Fc_{Low\_Desired}$.

$Fc_{Low\_Desired} = \sqrt{x^2 + x^2 + x^2 ...}$

$(Fc_{Low\_Desired})^2 = x^2 + x^2 + x^2 ...$

$(Fc_{Low\_Desired})^2 = (NumberOfCaps)x^2$

$x^2 = \frac{(Fc_{Low\_Desired})^2}{NumberOfCaps}$

$$xhz = \sqrt{\frac{(Fc_{Low\_Desired})^2}{NumberOfCaps}}$$

Calculate the capacitance value for each capacitor using its $R_{Thevinin}$ and the previously calculated Xhz for the desired Frequency Critical Low.

$$C_X = \frac{1}{2\pi(Rthev_{Cx})(xhz)}(farads)$$

Week 4

Multi-Stage Amplifier:
High-Frequency Response & Peaking

4.1 Objectives:

Multi-Stage Amplifier Design and Analysis:

High Frequency Response

- High Critical Frequency Fc_{High}: Understand and calculate the high critical frequency Fc_{High} for a multi-stage amplifier, taking into account the internal device capacitance and parasitic circuit elements in the amplifier stages.

Emitter Peaking

- Investigate the theory of Emitter Peaking.
- Calculate the Emitter Peaking capacitor and the Improvement Factor for a given circuit.

Shunt Peaking

- Investigate the theory of Shunt Peaking.
- Calculate the Shunt Peaking Inductor and the Improvement Factor for a given circuit.

Series Peaking

- Investigate the theory of Series Peaking.
- Calculate the Series Peaking Inductor and the Improvement Factor for a given circuit.

By achieving these objectives, students will develop a comprehensive understanding of an amplifier's High-Frequency Response and how to modify a circuit to improve frequency response using Emitter, Shunt, and Series Peaking.

4.2 Frequency Critical High:

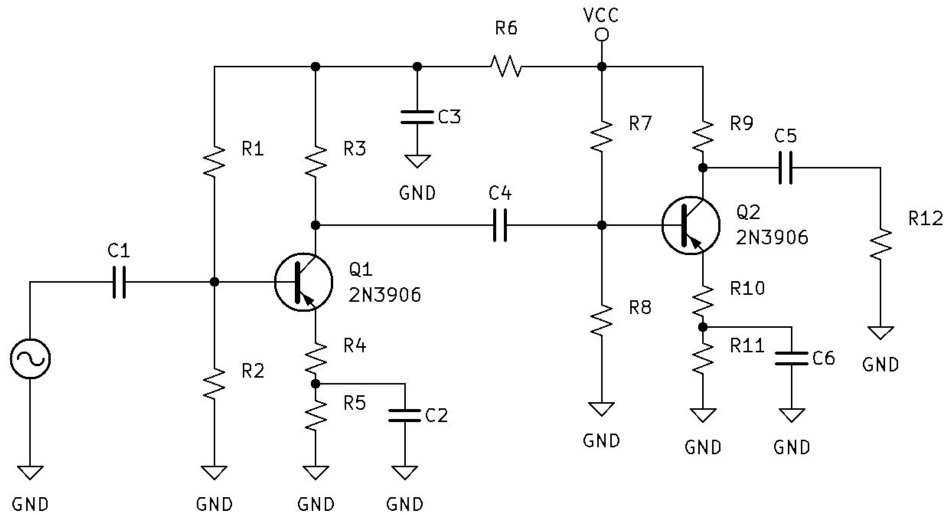


Figure 4.1: Two Stage Amplifier

4.2.1 Frequency Critical High Formula

For the two-stage amplifier, Frequency Critical High $FcHigh$ can be found using the following formula:

$$Fc_{High} = \frac{0.35}{\sqrt{(\frac{0.35}{FcH_{In}})^2+(\frac{0.35}{FcH_{Mid}})^2+(\frac{0.35}{FcH_{Out}})^2}}$$

SMALL–SIGNAL CHARACTERISTICS

Current–Gain – Bandwidth Product	(I_C = 10 mAdc, V_{CE} = 20 Vdc, f = 100 MHz)	f_T	250	–	MHz
Output Capacitance	(V_{CB} = 5.0 Vdc, I_E = 0, f = 1.0 MHz)	C_{obo}	–	4.5	pF
Input Capacitance	(V_{EB} = 0.5 Vdc, I_C = 0, f = 1.0 MHz)	C_{ibo}	–	10	pF
Input Impedance	(I_C = 1.0 mAdc, V_{CE} = 10 Vdc, f = 1.0 kHz)	h_{ie}	2.0	12	kΩ
Voltage Feedback Ratio	(I_C = 1.0 mAdc, V_{CE} = 10 Vdc, f = 1.0 kHz)	h_{re}	0.1	10	X 10^{-4}
Small–Signal Current Gain	(I_C = 1.0 mAdc, V_{CE} = 10 Vdc, f = 1.0 kHz)	h_{fe}	100	400	–
Output Admittance	(I_C = 1.0 mAdc, V_{CE} = 10 Vdc, f = 1.0 kHz)	h_{oe}	3.0	60	μmhos
Noise Figure	(I_C = 100 μAdc, V_{CE} = 5.0 Vdc, R_S = 1.0 kΩ, f = 1.0 kHz)	NF	–	4.0	dB

Figure 4.2: ON Semiconductor 2n3906 Data Sheet Excerpt

4.2.2 Frequency Critical High In

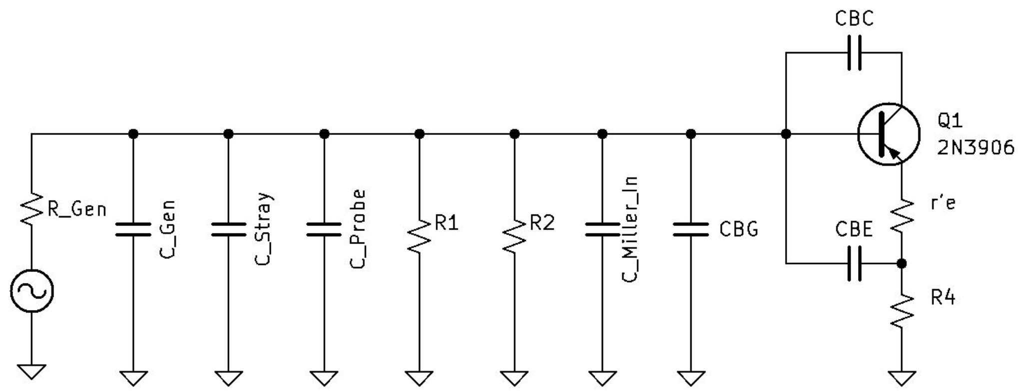


Figure 4.3: Input AC equivalent circuit at High Frequencies

Calculate the total capacitance at the input of the amplifier.

$$Ctotal_{IN} = C_{Gen} + C_{Stray} + C_{Probe} + C_{MillerIN} + C_{BG}$$

$$C_{Gen} = Specification\ in\ Manual$$
$$C_{Stray} \approx 10pF$$
$$C_{Probe} \approx 16pF, Specification\ in\ Manual$$
$$C_{MillerIN} = C_{OBO}(1 + \Delta V_{CE(Q1)})$$
$$C_{OBO} = transistor\ datasheet\ specification$$
$$\Delta V_{CE} = \frac{Vout}{Vin} = \frac{ic(RC//RL)}{ie(r'e+RE)} = \alpha\frac{RC//RL}{r'e+RE} = \alpha\frac{R3//RL_{eq}}{r'e+R4}$$
$$C_{BG} = C_{BE}(1 - \Delta V_{CC(Q1)})$$
$$C_{BE} = \frac{1}{2\pi f_{\tau} r'e}$$
$$f_{\tau} = transistor\ datasheet\ specification$$
$$\Delta V_{CC} = \frac{RE}{r'e+RE} = \frac{R4}{r'e+R4}$$

Calculate the Thevenin resistance for C_Total_{In}.

$$R\_Thevenin_{C\_TotalIn} = (R4 + r'e)(\beta + 1)//R2//R1//R_{Gen}$$

$$R\_Thevenin_{C\_TotalIn} \approx R_{Gen}$$

Calculate the Input High Critical Frequency.

$$FcH_{In} = \frac{1}{2\pi \times R\_Thev_{C\_TotalIn} \times C\_Total_{IN}}$$

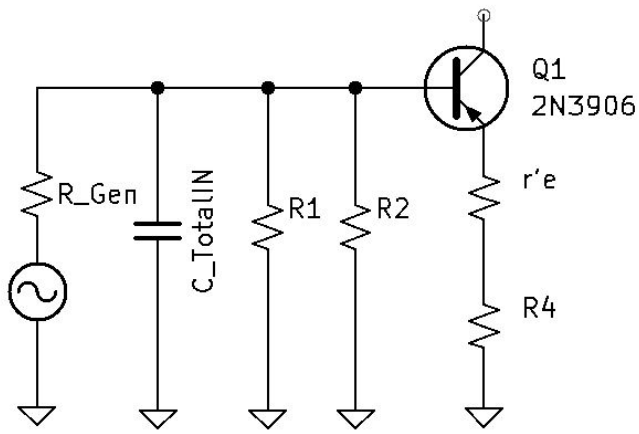


Figure 4.4: Capacitance Total In equivalent circuit at High Frequencies

4.2.3 Frequency Critical High Middle

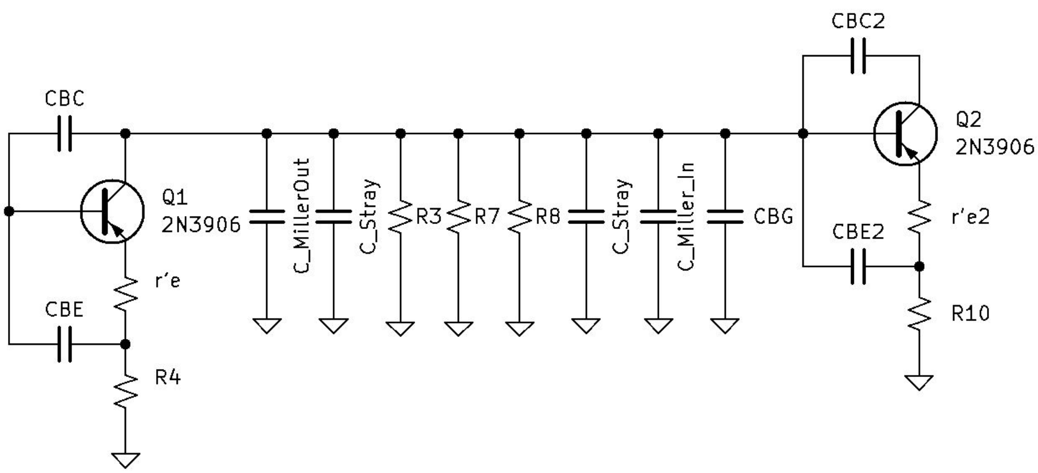


Figure 4.5: Middle AC equivalent circuit at High Frequencies

Calculate the total capacitance at the middle stage of the amplifier.

$$Ctotal_{Mid} = C_{MillerOut} + C_{Stray} + C_{Stray} + C_{MillerIN} + C_{BG}$$

$$C_{MillerOut} = CBC(\tfrac{1+\Delta V_{CE(Q1)}}{\Delta V_{CE(Q1)}})$$

$$CBC \approx C_{OBO}(Data\ Sheet)$$

$$C_{Stray} \approx 10pF$$

$$C_{MillerIN} = C_{OBO}(1 + \Delta V_{CE(Q2)})$$

$$C_{OBO} = transistor\ datasheet\ specification$$
$$\Delta V_{CE(Q2)} = \frac{Vout}{Vin} = \frac{ic(RC//RL)}{ie(r'e+RE)} = \alpha\frac{RC//RL}{r'e+RE} = \alpha\frac{R9//R12}{r'e+R10}$$
$$C_{BG} = C_{BE}(1 - \Delta V_{CC})$$
$$C_{BE} = \frac{1}{2\pi f_\tau r'e}$$
$$f_\tau = transistor\ datasheet\ specification$$
$$\Delta V_{CC(Q2)} = \frac{RE}{r'e+RE} = \frac{R10}{r'e+R10}$$

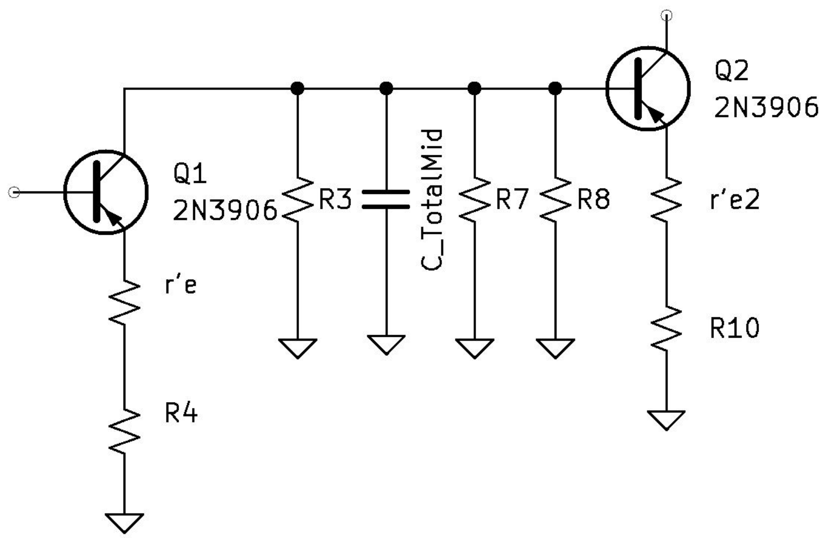


Figure 4.6: Capacitance Total Middle AC equivalent circuit at High Frequencies

Calculate the Thevenin resistance for C_Total_{Mid}.

$$R\_Thevenin_{C\_TotalMid} = (R10 + r'e2)(\beta + 1)//R3//R7//R8$$

Calculate the Middle High Critical Frequency.

$$FcH_{Mid} = \frac{1}{2\pi \times R\_Thev_{C\_TotalMid} \times C\_Total_{Mid}}$$

4.2.4 Frequency Critical High Out

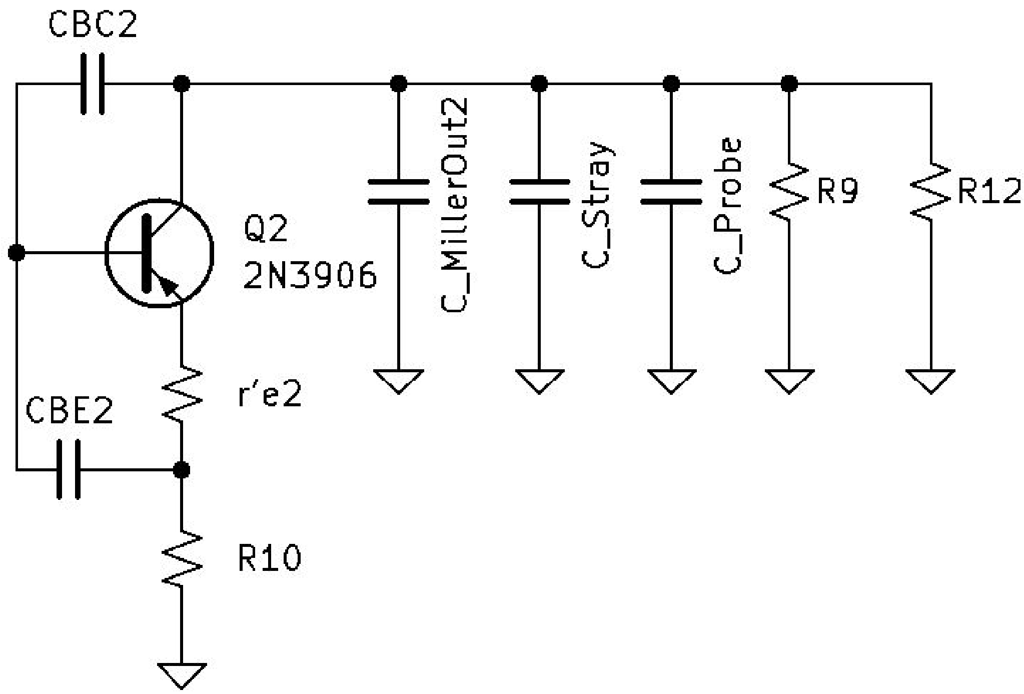


Figure 4.7: Output AC equivalent circuit at High Frequencies

Calculate the total capacitance at the output stage of the amplifier.

$$Ctotal_{Out} = C_{MillerOut2} + C_{Stray} + C_{Probe}$$

$$C_{MillerOut2} = CBC(\frac{1+\Delta V_{CE(Q2)}}{\Delta V_{CE(Q2)}})$$

$$CBC \approx C_{OBO}(Data\ Sheet)$$

$$C_{Stray} \approx 10pF$$

$$C_{Probe} \approx 16pF, Specification\ in\ Manual$$

Calculate the Thevenin resistance for C_Total_{Out}.

$$R\_Thevenin_{C\_TotalOut} = R9//R12$$

Calculate the Out High Critical Frequency.

$$FcH_{Out} = \frac{1}{2\pi \times R\_Thev_{C\_TotalOut} \times C\_Total_{Out}}$$

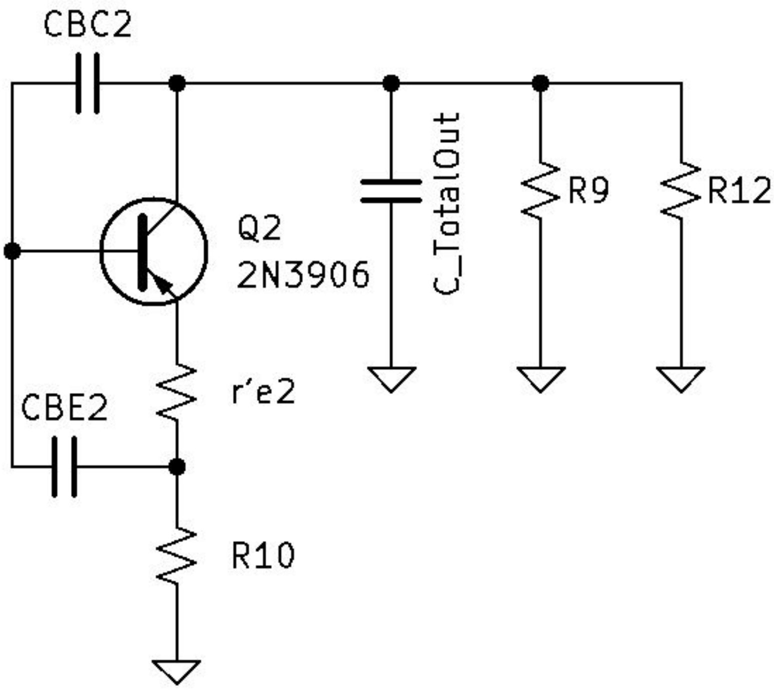


Figure 4.8: Capacitance Total Out AC equivalent circuit at High Frequencies

4.2.5 Frequency Critical High final calculation

Calculate the High Critical Frequency for the circuit using the previously calculated FcH_{In}, FcH_{Mid}, and FcH_{Out}.

$$Fc_{High} = \frac{0.35}{\sqrt{(\frac{0.35}{FcH_{In}})^2 + (\frac{0.35}{FcH_{Mid}})^2 + (\frac{0.35}{FcH_{Out}})^2}}$$

4.3 Emitter Peaking

4.3.1 Voltage Gain at Frequency Critical High

See figure 4.8 from the previous section. If we evaluate the ΔV formula we can see that at the FcH the numerator of the formula begins to roll off due to the parallel capacitive reactance.

$$\Delta V = \frac{Vout}{Vin}$$

$$\Delta V_{MidBand} = \alpha \frac{RL//RC}{RE+r'3} = \alpha \frac{R9//R12}{R10+r'e2}$$

$$\Delta V_{FcHigh} = \alpha \frac{R9//R12//XC_{C\_Out}}{R10+r'e2}$$

4.3.2 Emitter Peaking Explained

Knowing that the numerator is rolling off at a 20db/decade rate at the High Critical Frequency and that delta-V gap or difference is shrinking between the output voltage and the input voltage, how can we extend or prolong the difference between the output and the input?

One way is to apply Emitter Peaking. If the numerator is rolling off at FcHigh, can we also roll off the denominator of the formula at the same time? As both the numerator and the denominator roll the difference between them stays the same maintaining the gain. This can be achieved by placing an Emitter Peaking Capacitor in parallel R10 in our formula.

4.3.3 Solving for Emitter Peaking Capacitance

$$\Delta V_{FcHigh} = \alpha \frac{R9//R12//XC_{C\_Out}}{(XC_{EP}//R10)+r'e2}$$

Solve for the Emitter Peaking Capacitance value:

$$XC_{EP} = R10 \text{ at } Fc_{High}$$

$$C_{EP} = \frac{1}{2\pi \times R10 \times Fc_{High}}$$

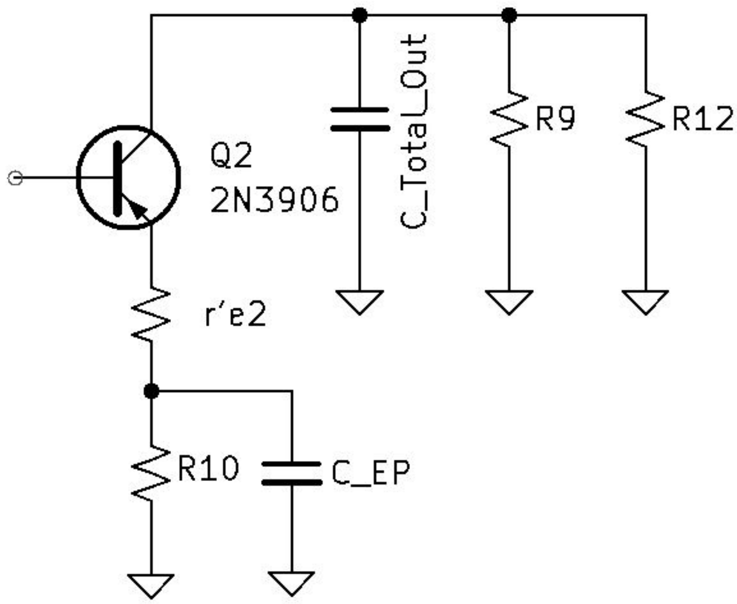


Figure 4.9: Emitter Peaking AC equivalent circuit

4.3.4 Improvement Factor (K)

Because Emitter Peaking involves RC circuits, we know that the roll-off will be 20dB/decade. If the old Fc_{High} was at -3dB and is now at 0dB with Emitter Peaking, it is reasonable to predict that the old -6dB frequency is now the new $FcHigh\_EP$.

Solving for the Emitter Peaking Improvement Factor (K)

$-6dB = 20Log\frac{1}{\sqrt{1+(\frac{F}{Fc_{High}})^2}}$ (standard RC Roll-Off dB formula set to -6dB)

$-6dB = 20Log\frac{1}{\sqrt{1+(\frac{K_{EP}}{1})^2}}$ (By setting Fc_{High} to 1, F becomes the Improvement Factor)

Solve for The Improvement Factor K_{EP}:

$$-6dB = 20Log\frac{1}{\sqrt{1+(\frac{K_{EP}}{1})^2}}$$

$$\frac{-6}{20} = Log\frac{1}{\sqrt{1+(\frac{K_{EP}}{1})^2}}$$

$$10^{\frac{-6}{20}} = \frac{1}{\sqrt{1+(\frac{K_{EP}}{1})^2}}$$

$$\sqrt{1 + (\tfrac{K_{EP}}{1})^2} = \frac{1}{10^{\frac{-6}{20}}}$$

$$1 + (\tfrac{K_{EP}}{1})^2 = (\frac{1}{10^{\frac{-6}{20}}})^2$$

$$(\tfrac{K_{EP}}{1})^2 = (\frac{1}{10^{\frac{-6}{20}}})^2 - 1$$

$$K_{EP} = \sqrt{(\frac{1}{10^{\frac{-6}{20}}})^2 - 1}$$

✓ $\boxed{K_{EP} = 1.727}$ (The Improvement Factor for Emitter Peaking is 1.727)

$$FcH_{EP} = K_{EP} \times Fc_{High}$$

✓ $\boxed{FcH_{EP} = 1.727 \times Fc_{High}}$

4.4 Shunt Peaking

4.4.1 Voltage Gain at Frequency Critical High

See figure 4.8 from the previous section. If we evaluate the ΔV formula we can see that at the FcH the numerator of the formula begins to roll off due to the parallel capacitive reactance.

$$\Delta V = \frac{Vout}{Vin}$$

$$\Delta V_{MidBand} = \alpha \frac{RL//RC}{RE+r'3} = \alpha \frac{R9//R12}{R10+r'e2}$$

$$\Delta V_{FcHigh} = \alpha \frac{R9//R12//XC_{C\_Out}}{R10+r'e2}$$

4.4.2 Shunt Peaking Explained

Emitter Peaking extends the High Critical Frequency by rolling off the denominator of the gain formula at the same time that the numerator is rolling off. Shunt Peaking attempts to prevent or delay the roll-off of the numerator. This is achieved by placing an inductor in series with the RC resistance of the final transistor stage. See Figure 4.10.

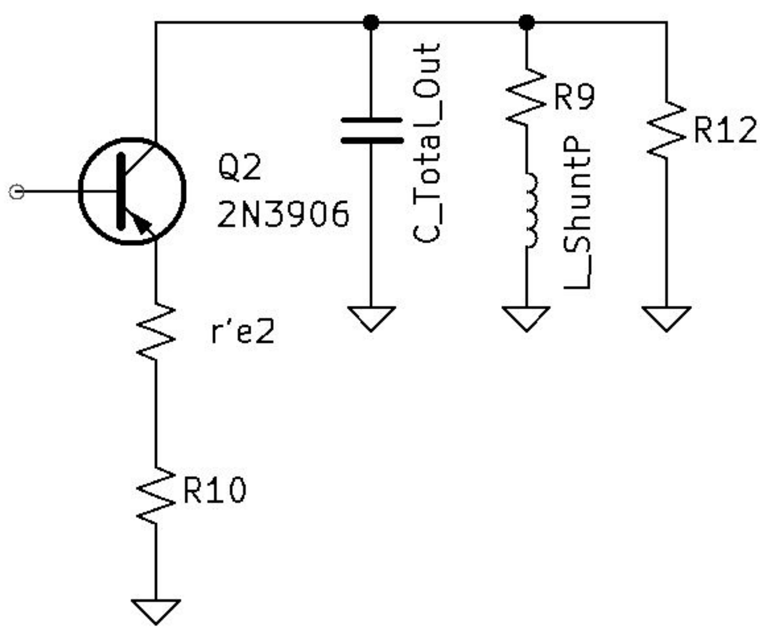


Figure 4.10: Shunt Peaking AC equivalent circuit

4.4.3 Improvement Factor (K)

For calculation purposes, design all Shunt Peaking with an improvement factor of 1.414.

✓ $\boxed{K_{Shunt} = 1.414}$

$$FcH_{ShuntP} = K_{Shunt} \times Fc_{High}$$

✓ $\boxed{FcH_{ShuntP} = 1.414 \times Fc_{High}}$

4.4.4 Solving for Shunt Peaking Inductance

For optimal flatness, make the frequency resonance 1.414 times larger than the High Critical Frequency of the circuit.

$$F_{Resonant} = Fc_{High} \times 1.414$$

$$F_{Resonant} = \frac{1}{2\pi\sqrt{LC}}$$

$$\sqrt{LC} = \frac{1}{2\pi \times F_{Resonant}}$$

$$LC = \frac{1}{(2\pi \times F_{Resonant})^2}$$

$$L = \frac{1}{C(2\pi \times F_{Resonant})^2}$$

✓ $\boxed{L = \dfrac{1}{C(2\pi \times Fc_{High} \times 1.414)^2}}$

4.5 Series Peaking

4.5.1 Voltage Gain at Frequency Critical High

See figure 4.8 from the previous section. If we evaluate the ΔV formula we can see that at the FcH the numerator of the formula begins to roll off due to the parallel capacitive reactance.

$$\Delta V = \frac{Vout}{Vin}$$

$$\Delta V_{MidBand} = \alpha \frac{RL//RC}{RE+r'3} = \alpha \frac{R9//R12}{R10+r'e2}$$

$$\Delta V_{FcHigh} = \alpha \frac{R9//R12//XC_{C\_Out}}{R10+r'e2}$$

4.5.2 Series Peaking Explained

If Emitter Peaking extends the High Critical Frequency by rolling off the denominator of the gain formula at the same time that the numerator is rolling off and Shunt Peaking attempts to prevent or delay the roll-off of the numerator of the gain formula. Then you can think of Series Peaking as independent of the gain formula. Series peaking relies on the concept of LC circuit voltage magnification, where at resonance, the capacitive reactance will appear to have more voltage than the source. This is achieved by placing an inductor in series with a load. See Figure 4.12.

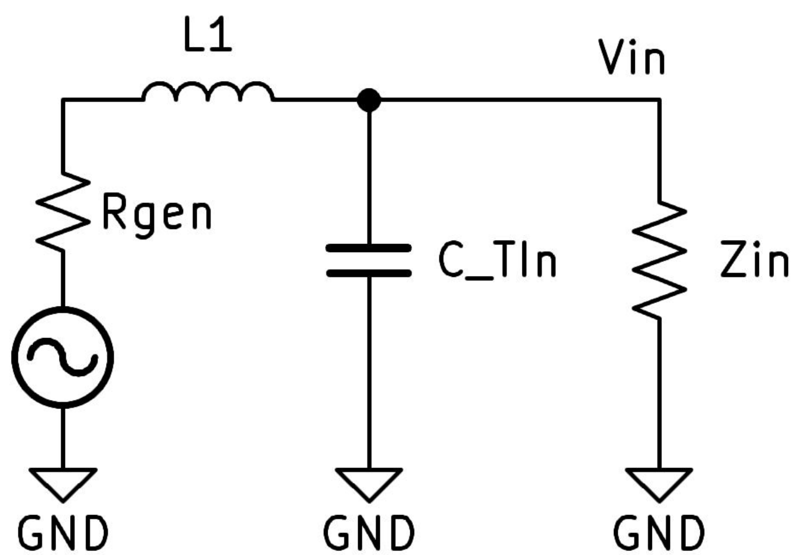


Figure 4.11: Series Peaking AC equivalent circuit

4.5.3 Improvement Factor (K)

For calculation purposes, we will design all Series Peaking with an improvement factor of 1.414.

✓ $\boxed{K_{Series} = 1.414}$

$FcH_{SeriesP} = K_{Series} \times Fc_{High}$

✓ $\boxed{FcH_{SeriesP} = 1.414 \times Fc_{High}}$

4.5.4 Solving for Series Peaking Inductance

For optimal flatness, make X_L 1.414 times smaller than X_C at the High Critical Frequency of the circuit.

$X_L = \frac{X_C}{1.414}$

$X_L = X_C \times 0.707$ (multiplying by 0.707 is the same as dividing by 1.414)

$X_L = \frac{1}{2\pi C_{Tin} FC_{High}} \times 0.707$ (substitute X_C = formula)

$X_L = 2\pi FL$ (standard X_L formula)

$X_L = 2\pi FC_{High} L$ (substitute FC_{High} for F)

$\frac{1}{2\pi C_{Tin} FC_{High}} \times 0.707 = 2\pi FC_{High} L$ (substitute X_L)

$\frac{1}{C_{Tin}(2\pi FC_{High})^2} \times 0.707 = L$ (solve for L)

✓ $\boxed{L = \frac{0.707}{C_{Tin}(2\pi FC_{High})^2}}$ (cleanup formula)

4.5.5 Alternative Method Series Peaking

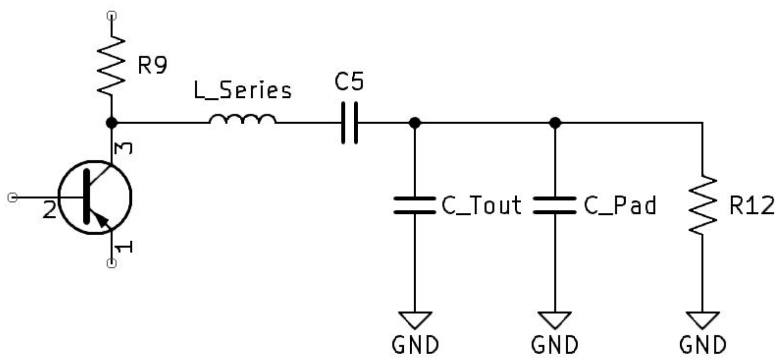


Figure 4.12: Series Peaking AC equivalent circuit

The improvement factor (K) for the alternative Series Peaking method is the same as the traditional method, see section 4.5.3.

The Series Inductor for the alternative Series Peaking method can be found using the same formula as section 4.5.4, however the total capacitance out will need to be substituted for capacitance total in.

$$✓\ \boxed{L = \frac{0.707}{(C_{Tout} + C_{pad})(2\pi F C_{High})^2}}$$

4.6 Factors that affect Frequency Response

Factors that affect Low Critical Frequency Fc_{Low}

- Coupling Capacitors & Bypass Capacitors
- Decoupling Networks
- Power Supply Filters
- Resistance Values

Factors that affect High Critical Frequency Fc_{High}

- Device Capacitance
- Stray Capacitance
- Generator Capacitance
- Gain
- $f\tau$ & Slew Rate
- Resistance Values
- Probe Capacitance

4.7 Types of Amplifier Distortion

Amplitude Distortion

- **Amplitude Distortion** is defined as the inability of an amplifier to reproduce an output that is a linear function of the input. Clipping and Cross-Over Distortion are types of amplitude distortion.
- In high-fidelity audio systems, minimizing amplitude distortion is crucial to maintaining the accuracy and faithfulness of the reproduced sound. Engineers and designers aim to create amplifiers with low levels of amplitude distortion to ensure that the output closely matches the input signal, preserving the integrity of the audio source.

Frequency Distortion

- **Frequency Distortion** is defined as the inability of an amplifier to amplify all of the desired frequencies with the same gain.
- Measuring frequency distortion often involves analyzing the system's frequency response. Engineers aim to design systems with flat and linear frequency responses to minimize frequency distortion and ensure accurate reproduction of the input signal.

Phase Distortion

- **Phase Distortion** is defined as the inability of desired frequencies with the same time delay.
- Phase distortion is undesirable in stereo systems, and careful attention is given to signal polarity as well as speaker placement to ensure accurate phase coherence and faithful audio reproduction.

Crossover Distortion

- **Crossover Distortion** occurs in a Push-Pull, Class AB amplifier when both transistors are off.
- Crossover distortion is undesirable in audio amplification systems because it can introduce odd-order harmonics into the signal, leading to a harsh and distorted sound. To mitigate crossover distortion, amplifier designers often use techniques like biasing, which involves applying a small DC voltage to the transistors to keep them slightly conducting even when there is no input signal, reducing the gap between their active regions and minimizing crossover distortion.

Week 5a

Amplifier Classifications

5a.1 Objectives

Amplifier Classifications:

Define the different types of amplifier classifications.

- Identify the following for each amplifier classification:
  - Linearity and fidelity qualities
  - Efficiency characteristics
  - Typical applications

5a.2 Amplifier Classifications

Class A Amplifiers

- The Universal Biased Common Emitter Amplifier is an example of a Class A amplifier.
- Transistor Amplification is active for 360° of the input cycle (always on).
- Achieves the highest degree of linearity.
- Low Efficiency, $MaxEffiency_{ClassA} \approx 25\%$

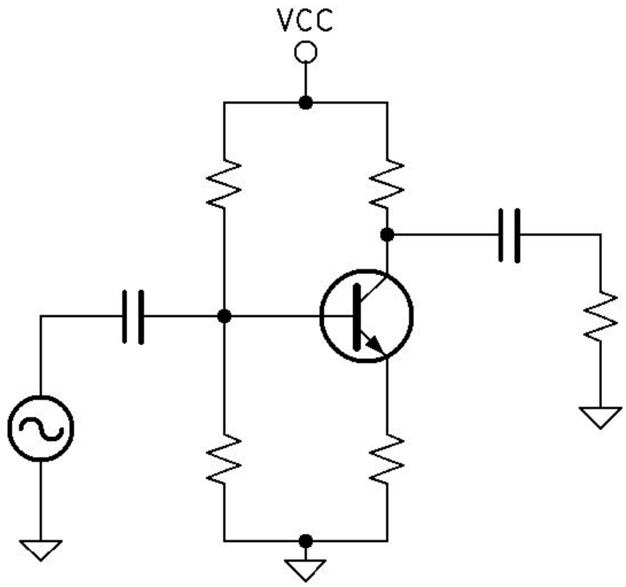


Figure 5a.1: Class A Amplifier

Class B Amplifiers

- Transistor Amplification is active for 180° of the input cycle (on half the time).
- Achieves a moderate degree of linearity.
- moderate Efficiency, $MaxEffiency_{ClassA} \approx 50\%$

Class AB Amplifiers

- Also known as Push-Pull.
- Transistor Amplification is active for each transistor nearly 180° of the input cycle. This means that nearly 360° of the input signal is amplified.
- Crossover distortion occurs when both transistors are off.
- Achieves a high degree of linearity.
- moderate Efficiency, $MaxEffiency_{ClassA} \approx 50\%$

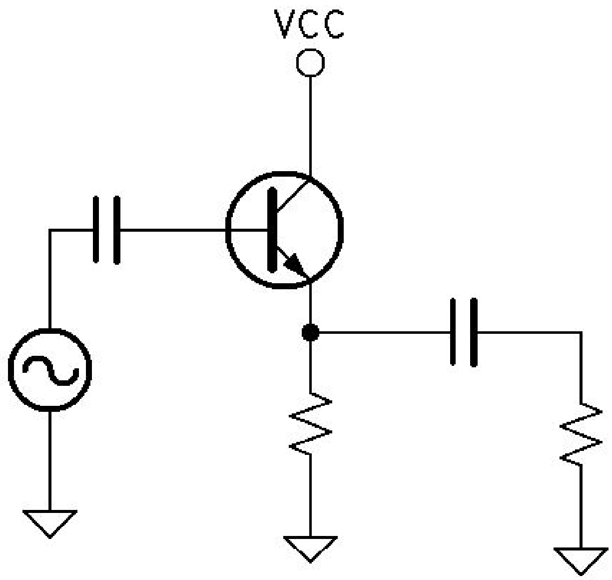


Figure 5a.2: Class B Amplifier

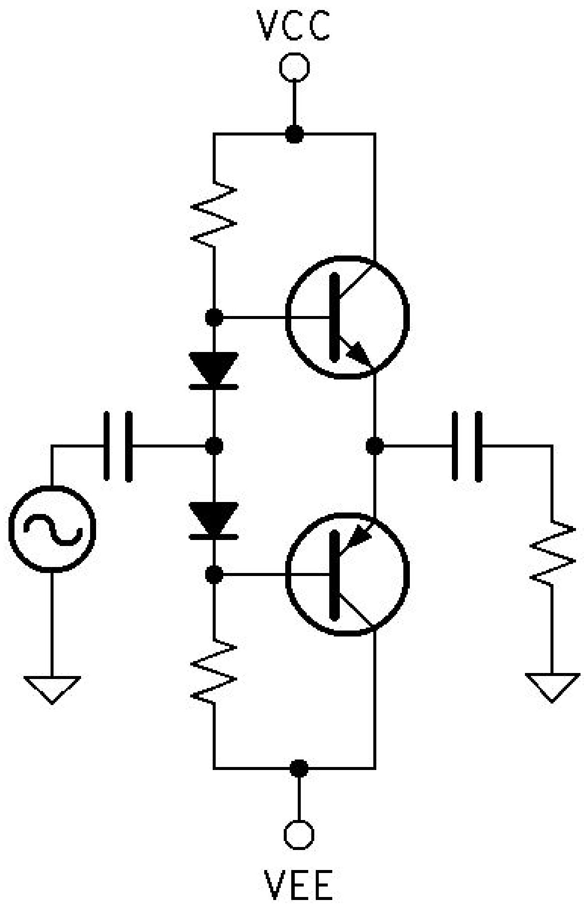


Figure 5a.3: Class AB Amplifier

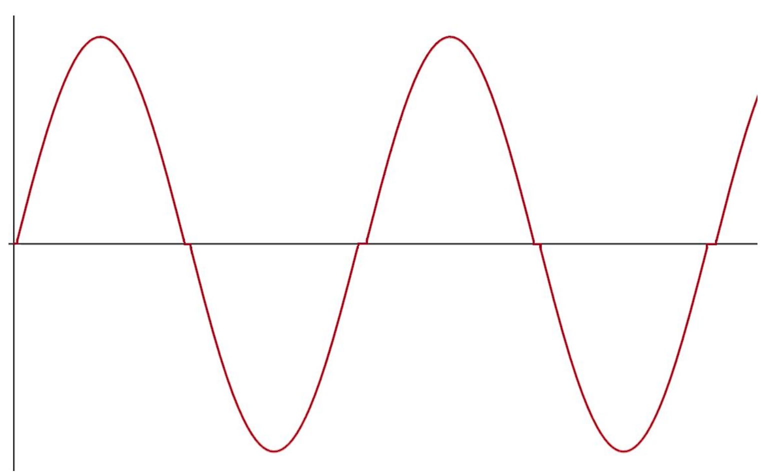


Figure 5a.4: Class AB, Crossover Distortion

Class C Amplifiers

- Transistor Amplification is active less than 180° of the input cycle (less than 50% of the time).
- Common uses high-frequency sine-wave oscillators & radio frequency amplifiers.
- Achieves a low degree of linearity.
- moderate Efficiency, $MaxEffiency_{ClassA} \approx 75\%$

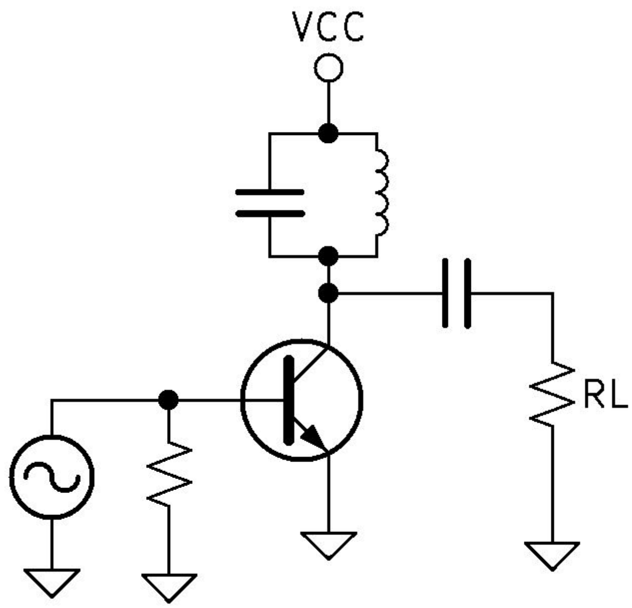


Figure 5a.5: Class C Amplifier

Class D Amplifiers

- Non-linear switching amplifier also known as a PWM amplifier.
- Voltage and current waveforms do not overlap. The transistor is operating in either the Saturation or Cutoff region.
- Can achieve a high degree of linearity using an additional integration circuit that converts the PWM signal back to intelligence.
- Highly Efficient, $MaxEffiency_{ClassA} \approx 100\%$

Week 5b

Push-Pull Amplifiers

5b.1 Objectives

Push-Pull Amplifiers:

- Design and design considerations for Push-Pull amplifiers.
- Analysis of Push-Pull amplifiers to include Z_{IN}, Z_{OUT}, and gains.

5b.2 Push-Pull, Class AB Amplifier Characteristics:

- $\Delta V \approx 1$, and ΔI is high.
- Z_{IN} is high, and Z_{OUT} is low.
- Class AB amplifiers have improved efficiency when compared to Class A amplifiers.
- $MaxEfficiency_{ClassAB} \approx 50\%$, $MaxEfficiency_{ClassA} \approx 25\%$.
- Class AB amplifiers achieve a high degree of linearity.
- Class AB amplifiers are susceptible to Crossover Distortion which occurs when both transistors are off.

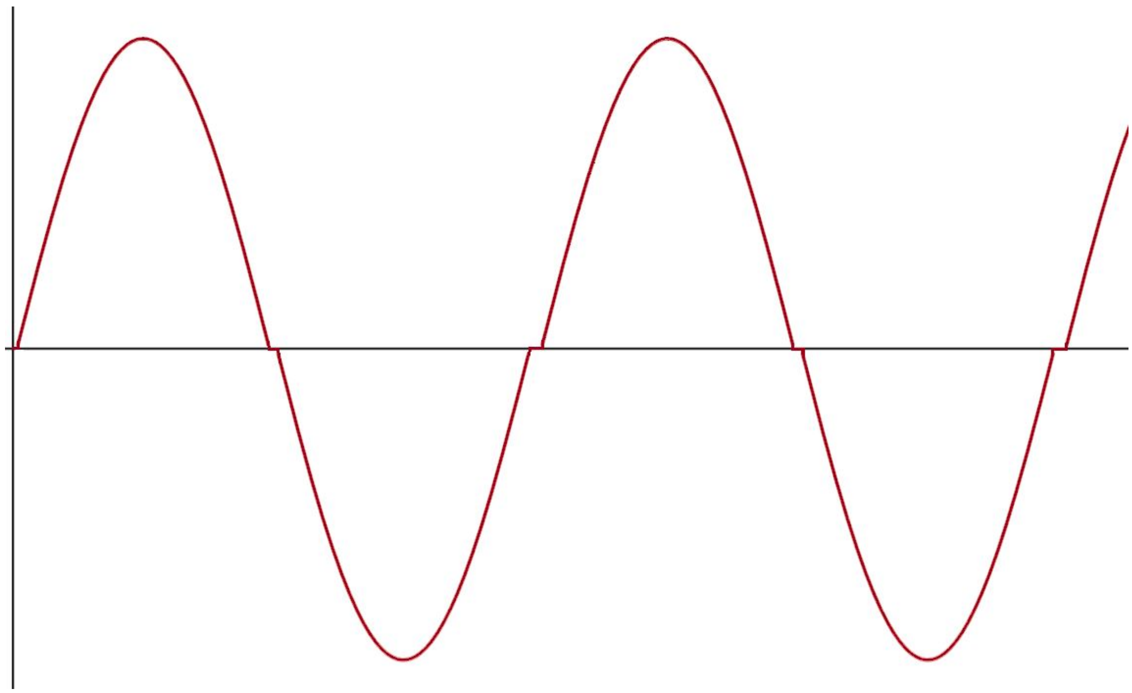


Figure 5b.1: Class AB, Crossover Distortion

5b.3 Push-Pull Amplifier Design

5b.3.1 Split Rail Voltage VCC & VEE

The main benefit when using split rail voltage is the potential to ground between the two transistors will be at zero volts allowing the output coupling capacitor C2 to be a relatively low voltage which reduces the capacitor's cost. High Voltage High Capacitance capacitors are expensive. Low Voltage High Capacitance capacitors are less expensive.

5b.3.2 Design steps and formulas:

$Q1_{\beta eta} \approx Q2_{\beta eta}$

$\Delta V \approx 1$

$Vin_{MaxP} \approx Vout_{MaxP} \approx \frac{VCC+VEE}{2}$

$Ie_{MaxP} = \frac{Vout_{MaxP}}{RL}$

$Ib_{MaxP} = \frac{Ie_{MaxP}}{\beta+1}$

$IR1 \geq 11 \times Ib_{maxP}$

$VR1 = \frac{VCC+VEE}{2} - V_{D1}$

$R1 = \frac{VR1}{IR1}$

$R2 = R1$

Z_{in}:(Analyze with Q1 on and Q2 off)

$Z_{in} = ((((RL + r'e_{Q1})(\beta + 1))//R1) + r'd1)//(r'd2 + R2)$

$r'd = \frac{0.026}{I_{D1}}$

$Z_{in} \approx R1//R2$

Z_{out} :(Analyze with Q1 on and Q2 off)

$Z_{out} = \frac{((R_{Gen}//(r'd2+R2))+r'd1)//R1}{\beta+1} + r'e1$

$Z_{out} \approx r'e1$

$PRL_{Max} \approx \frac{(0.707 \times Vout_{MaxP})^2}{RL}$

$PQ1_{Max} \approx \frac{PRL_{Max}}{2}$

$PQ2_{Max} = PQ1_{Max}$

5b.3.3 Improved Input Impedance Zin and Power Capabilities using the Darlington Pair

Using Darlington Pairs will improve both power capabilities and the input impedance Z_{in} of the amplifier. Q2 and Q4 can be power transistors like the TIP41 and TIP42. Power transistors typically have a lower beta than their low-power general-purpose counterparts. Design using the previous section steps while adjusting the formulas to account for the additional components in the circuit.

5b.3.4 Shoot-Through Protection using Swamping Resistors

Shoot-Through

Observe Figure 5b.4. A potentially serious problem with Push-Pull Amplifiers is Shoot-Through. Shoot-Through occurs when both power transistors (Q2 and Q4) turn on at the same time due to excessive bias voltage. This creates a path for current from VCC directly to VEE with little or no opposition (resistance). By placing Swamping Resistors R3 and R4 in the current path we can limit the Shoot-Through current.

Swamper Resistor Value

Typically, a Push-Pull Amplifier is used to provide high current gain for a small load resistance, for example, an 8Ω speaker. Considering our load resistance if we used 8Ω Swamping Resistors we would lose half of our signal voltage across at the Swampers. Ideally, the Swamper Resistors will be at a minimum 10x smaller than the load resistance.

$$R_{Swamper} \leq \frac{RL}{10}$$

$$R3 \leq \frac{RL}{10}$$

$$R4 = R3$$

5b.3.5 Crossover Distortion

Observe Figure 5b.5. Crossover distortion occurs when both power transistors are off at the same time and the input signal has not overcome the biasing needs to turn on one of the transistors. Observe Figure 5b.5. By removing two of the diodes, the Crossover Distortion should become observable.

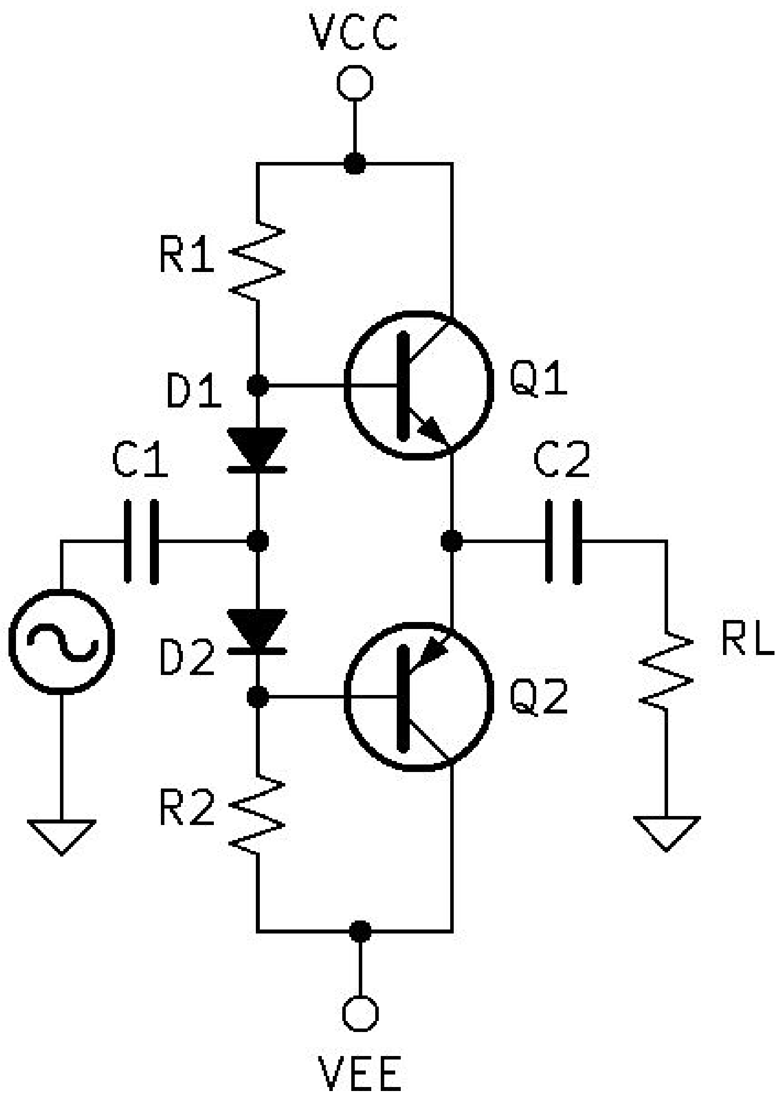


Figure 5b.2: Class AB Amplifier

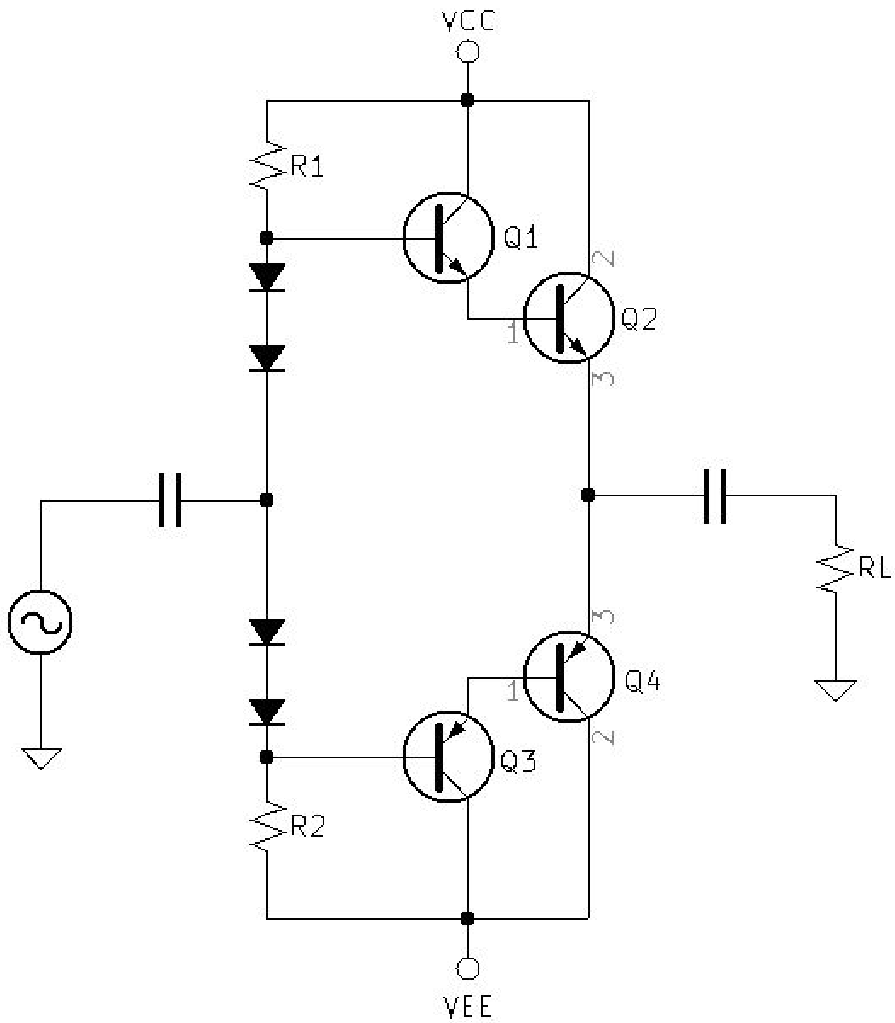


Figure 5b.3: Class AB Amplifier with Darlington Pair for High Power Applications

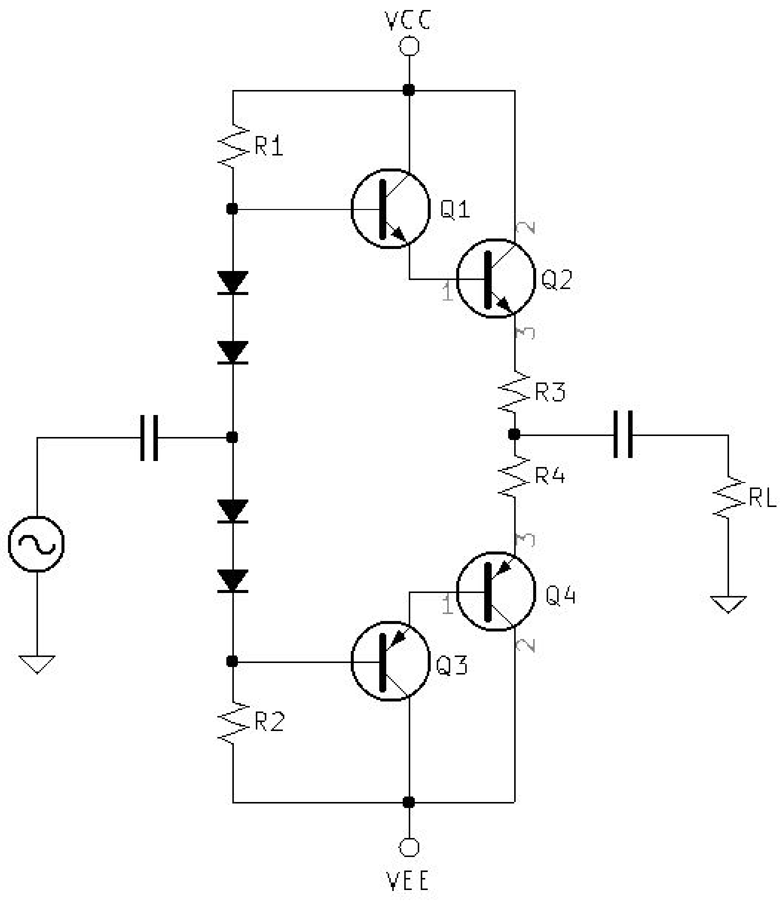


Figure 5b.4: Class AB Amplifier with Swamping Resistors R3 and R4

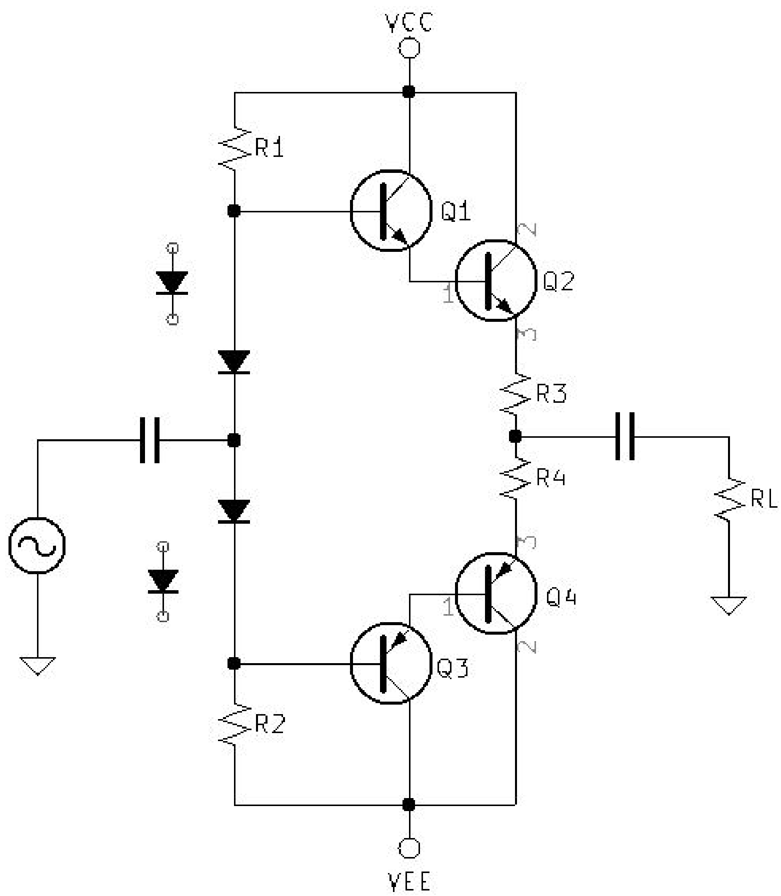


Figure 5b.5: Crossover Distortion

5b.3.6 Measuring Crossover Distortion with FFT

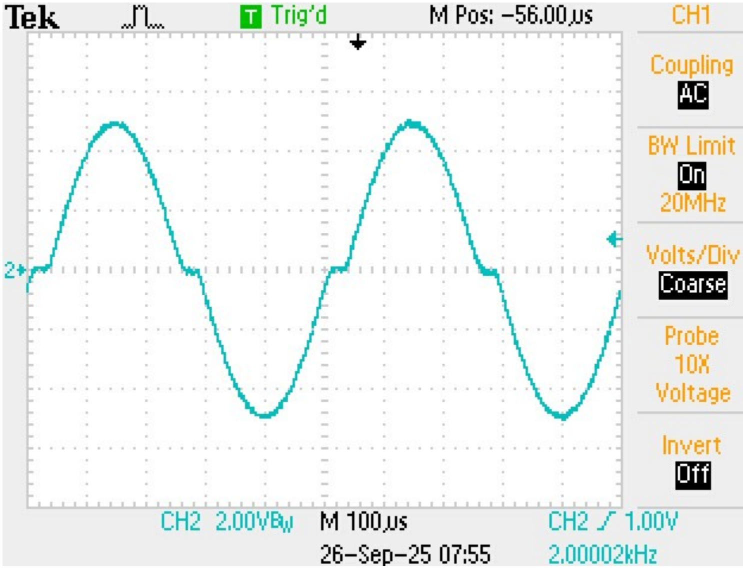


Figure 5b.6: Push-Pull Crossover Distortion as seen on Oscilloscope

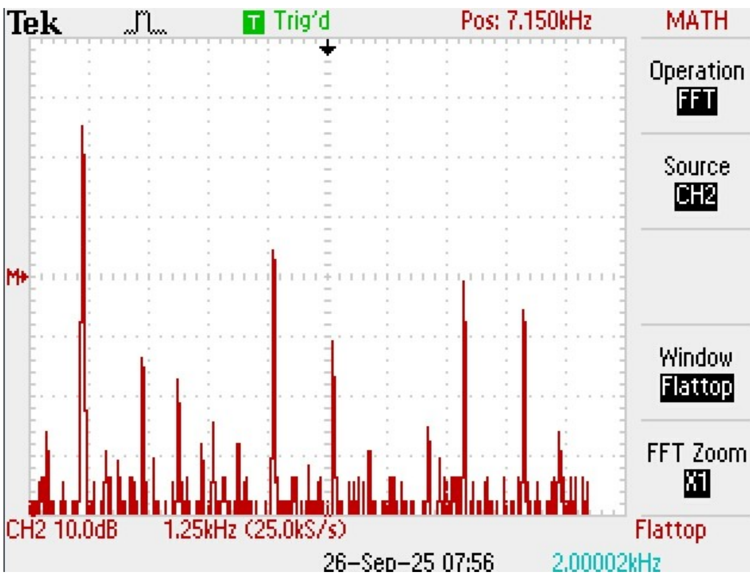


Figure 5b.7: Push-Pull Crossover Distortion FFT Measurement

5b.3.7 Push-Pull Calibration

Calibrating the circuit to have a small amount of acceptable Shoot-Through current will eliminate Crossover Distortion. This is accomplished by adjusting the bias voltage.

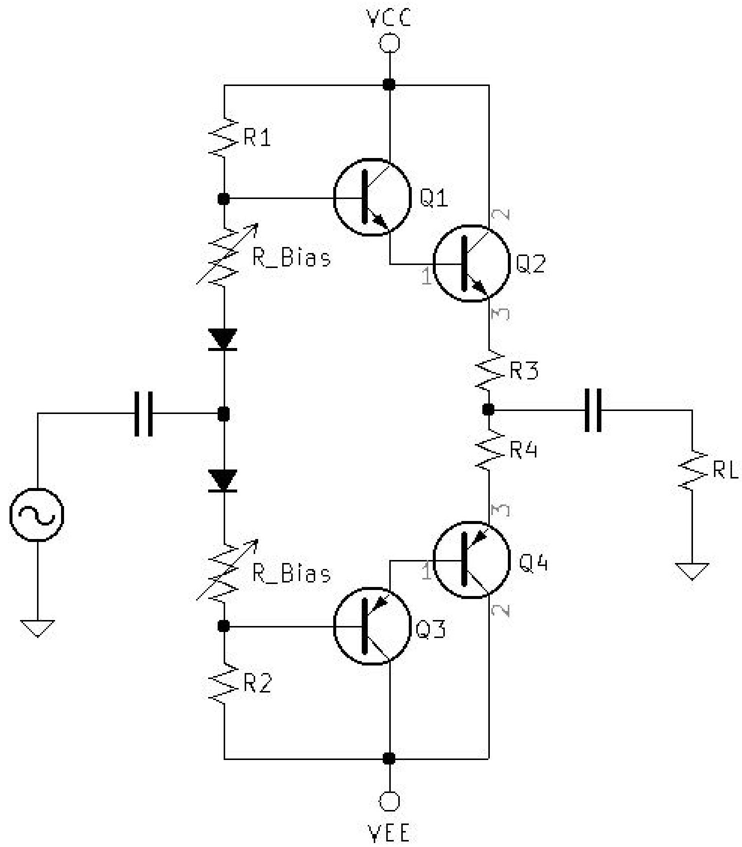


Figure 5b.8: Push-Pull Calibration

Bias Voltage Calibration & Quiescent Current

Output Stage Bias Current

The quiescent bias current I_q of the output stage plays a critical role in controlling crossover distortion. It is important that the right amount of bias current flows through the output stage, from top to bottom, when the output stage is not delivering any current to the load. Notice that together the two driver and two output transistors require at least four V_{be} voltage drops from the base of Q1 to the base of Q2 to begin to turn on. Any additional drop across the output emitter resistors will increase the required bias-spreading voltage.

The required bias voltage for the output stage is developed across the bias spreader, which is usually a V_{be} multiplier. The objective of the bias spreader design is temperature stability of the output stage quiescent current. The temperature coefficient of the voltage produced by the V_{be} multiplier should match that of the base-emitter junction voltages of the driver and output transistors. Since the V_{be} of a transistor decreases 2.2 mV/°C, it is important for thermal bias stability that these junction drops track one another. The output transistors will usually heat up the most. Because they are mounted on a heat sink, the V_{be} multiplier transistor is often mounted on the heat sink so that it is exposed to the same approximate temperature. This approach is only an approximation, because the drivers are often not mounted on the heat sink and the temperature of the heat sink changes more slowly than that of the power transistor junctions.

Figure 5b.9: Designing Audio Power Amplifiers Bob Cordell [12, p. 168]

Shoot-through current is an out-of-control current that can damage an amplifier. Quiescent current is a controlled bias current used to eliminate crossover distortion in push-pull amplifiers. A quiescent current of $\approx 30mA$ will eliminate Crossover Distortion. When testing the DC biasing of the designed Push-Pull circuit with the Swamping Resistors (Fig. 5b.4), one of three scenarios will occur.

Scenario One: The measured current through the Swamper resistors is approximately 30mA, indicating that the Push-Pull is properly calibrated.

✓ $I_{Quiescent} = \frac{V_{R3}}{R3} \approx 30mA$

Scenario Two: The measured current through the Swamper resistors is greater than 30mA, indicating that the Push-Pull may be over-biased and susceptible to shoot-through.

✗ $I_{Quiescent} = \frac{V_{R3}}{R3} > 30mA$

Reduce the bias voltage across the Swamping Resistors. Measure the voltage from the base of Q1 to the Emitter of Q2 then add the desired voltage for R3. Remove one diode per side as seen in Fig. 5b.8. Calculate the proper resistor value to achieve the desired voltage.

$R_{Bias} < \frac{V_{Diode}}{I_{Diode}}$

Scenario Three: The measured current through the Swamper resistors is less than 30mA indicating that the Push-Pull is under-biased, which could result in crossover distortion.

✗ $I_{Quiescent} = \frac{V_{R3}}{R3} < 30mA$

Increase the bias voltage across the Swamping Resistors. Measure the voltage from the base of Q1 to the Emitter of Q2 then add the desired voltage for R3. Remove one diode per side as seen in Fig. 5b.8. Calculate the proper resistor value to achieve the desired voltage.

$R_{Bias} > \frac{V_{Diode}}{I_{Diode}}$

Once bias calibration is complete and the quiescent current is approximately 30mA, the crossover distortion and shoot-through current should be eliminated.

5b.3.8 Push-Pull FFT Measurements

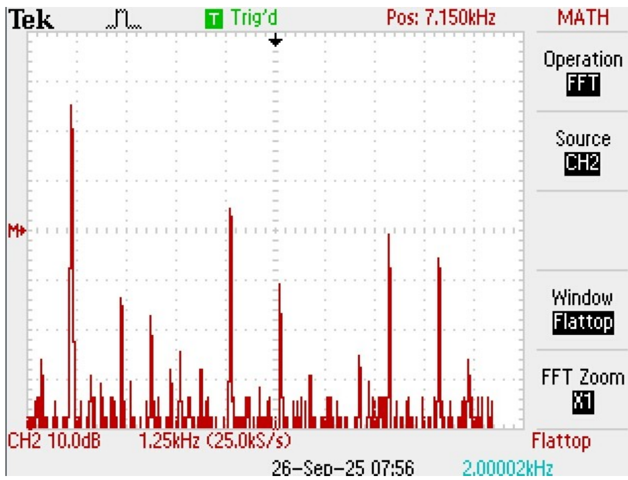


Figure 5b.10: Crossover Distortion FFT Measurement
(Sine-wave with visible Cross-Over when measured with the Oscilloscope)

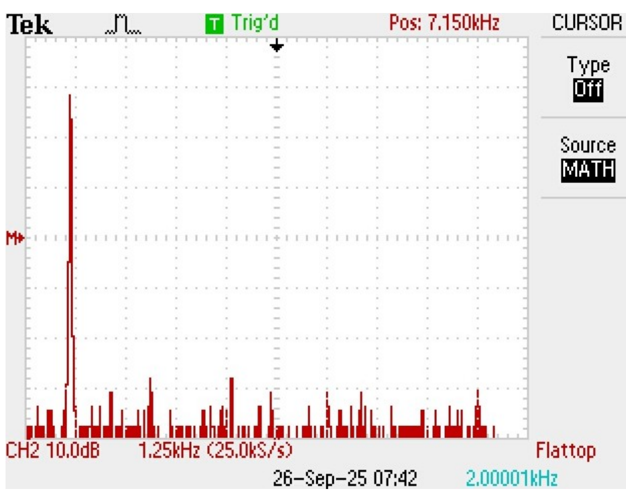


Figure 5b.11: Properly Biased Push Pull FFT Measurement
(Sine-wave with no visible crossover when measured with the Oscilloscope)

5b.3.9 Push-Pull Vout Max Peak Calculations

The Vout Max Peak of a Push-Pull Amplifier should be approximately equal to the VCE voltage of one of the power transistors, as long as the Swamping resistor is ten times smaller than the load resistance.

$$Vout_{MaxP} \approx VCE_{Q2}$$

$$Vout_{MaxP} = \frac{VCE_{Q2}}{R3+RL} \times RL$$

5b.3.10 Push-Pull Zin Calculations

Referencing Figure 5b.4. Zin is the impedance that the Generator sees.

$$Zin \approx R1//R2$$

Analyze as though Q1 is on and Q4 is off:

$$Zin = (((((RL+R3+r'e_{Q2}) \times (\beta_{Q2}+1)) + r'e_{Q1}) \times (\beta_{Q1}+1))//R1) + r'd + r'd)//(r'd + r'd + R2)$$

➥ $r'd = \frac{0.026v}{I_D}$

5b.3.11 Push-Pull Zout Calculations

Referencing Figure 5b.4. Zout is the impedance that the Load sees.

$$Zout \approx r'e_{Q2} + R3$$

Analyze as though Q1 is on and Q4 is off:

$$Zout = \frac{(\frac{((Rgen//(2r'd+R2))+2r'd)//R1}{\beta_{Q1}+1} + r'e_{Q1})}{\beta_{Q2}+1} + r'e_{Q2} + R3$$

➥ $r'd = \frac{0.026v}{I_D}$

Week 5c

Heat-Sinks

5c.1 Objectives

Heat-Sinks:

- Locate and identify the thermal resistance for a given heat sink using a data sheet.
- Calculate practical power usage for a given heat-sink circuit application using thermal resistance.

5c.2 Heat Sink Calculation examples:

5c.2.1 TIP41 Thermal Considerations

Operating and Storage Junction, Temperature Range	T_J, T_{stg}	−65 to +150	°C

Figure 5c.1: TIP41 Maximum Temperature data

The Maximum Operating Temperature for a TIP41 according to the data-sheet is 150°C. To avoid part damage, I like to design for a Maximum Design Temperature of 90% of the absolute maximum temperature.

$$MaxTemp_{Design} = 0.9(MaxTemp_{Absolute})$$

$$TIP41\_MaxTemp_{90\%} = 0.9(TIP41\_MaxTemp_{Absolute})$$

$$TIP41\_MaxTemp_{90\%} = 0.9(150°C)$$

$$TIP41\_MaxTemp_{90\%} = 135°C$$

Compensate for ambient temperature.

$$TIP41\_MaxTemp_{Design} = 135°C - Ambient$$
$$TIP41\_MaxTemp_{Design} = 135°C - 25°C$$
$$TIP41\_MaxTemp_{Design} = 110°C$$

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction–to–Case	$R_{\theta JC}$	1.67	°C/W
Thermal Resistance, Junction–to–Ambient	$R_{\theta JA}$	57	°C/W

Figure 5c.2: TIP41 Thermal Resistance data

The Junction-to-Ambient Thermal Resistance is used when the device has no heat sink attached.

$ThermalR_{Ambient} = \frac{57°}{1W}$

If the Maximum desired Temperature is 135 degrees C, Solve for the Maximum power in Watts.

$\frac{57°}{1W} = \frac{110°}{X}$

$X(57°) = (110°)1W$

$X = \frac{(110°)1W}{(57°)}$

$X = \frac{110W}{57}$ (degrees cancel)

$X = 1.93W$

$MaxPowerDissipation_{TIP\_Ambient} = 1.93W$ (No Heat Sink)

5c.2.2 HSE-B20254-035H

When using a heat sink, add the Thermal Resistance in series to find the Total Thermal Resistance between the device and the heat sink.

HSE-B20254-035H Thermal Resistance = $12.93°C/W$

TIP41 Thermal Resistance Junction to Case= $1.67°C/W$

Total Thermal Resistance = $12.93°C/W + 1.67°C/W$

SERIES: HSE-BX-02 | **DESCRIPTION:** HEAT SINK

FEATURES

- TO-220 package
- placement pins for secure PCB attachment
- round hole for component attachment
- multiple available cut lengths

RoHS

MODEL

	length	thermal resistance[1]				power dissipation[1]
		@ 75°C ΔT, nat conv (°C/W)	@ 1 W, nat conv (°C/W)	@ 1 W, 200 LFM (°C/W)	@ 1 W, 400 LFM (°C/W)	@ 75°C ΔT, nat conv (W)
	(mm)					
HSE-B20254-035H	25.4	12.93	14.40	3.28	2.49	5.80
HSE-B20381-035H	38.1	11.54	13.64	3.66	2.76	6.50
HSE-B20508-035H	50.8	9.62	12.98	5.17	3.28	7.80
HSE-B20508-035H-W[2]	50.8	9.62	12.98	5.17	3.28	7.80
HSE-B20635-035H	63.5	8.15	10.92	4.35	2.86	9.20
HSE-B20635-035H-W[2]	63.5	8.15	10.92	4.35	2.86	9.20

Figure 5c.3: HSE-BX-02 series Heat Sink data

Total Thermal Resistance $= 14.6°C/W$

Divide the Total Thermal Resistance into the Designed Maximum Temperature to find the Maximum Power for the HSE-B20254-035H.

HSE-B20254-035H$_{MaxPower} = \frac{MaxTemp_{Design}}{ThermalR_{Total}}$

HSE-B20254-035H$_{MaxPower} = \frac{110°C}{14.6°C/W}$

✓ HSE-B20254-035H$_{MaxPower} = 7.534\ Watts$

5c.2.3 HSE-B20635-035H

When using a heat sink, add the Thermal Resistance in series to find the Total Thermal Resistance between the device and the heat sink.

HSE-B20635-035H Thermal Resistance $= 8.15°C/W$

TIP41 Thermal Resistance Junction to Case$= 1.67°C/W$

$$\text{Total Thermal Resistance} = 8.15°C/W + 1.67°C/W$$

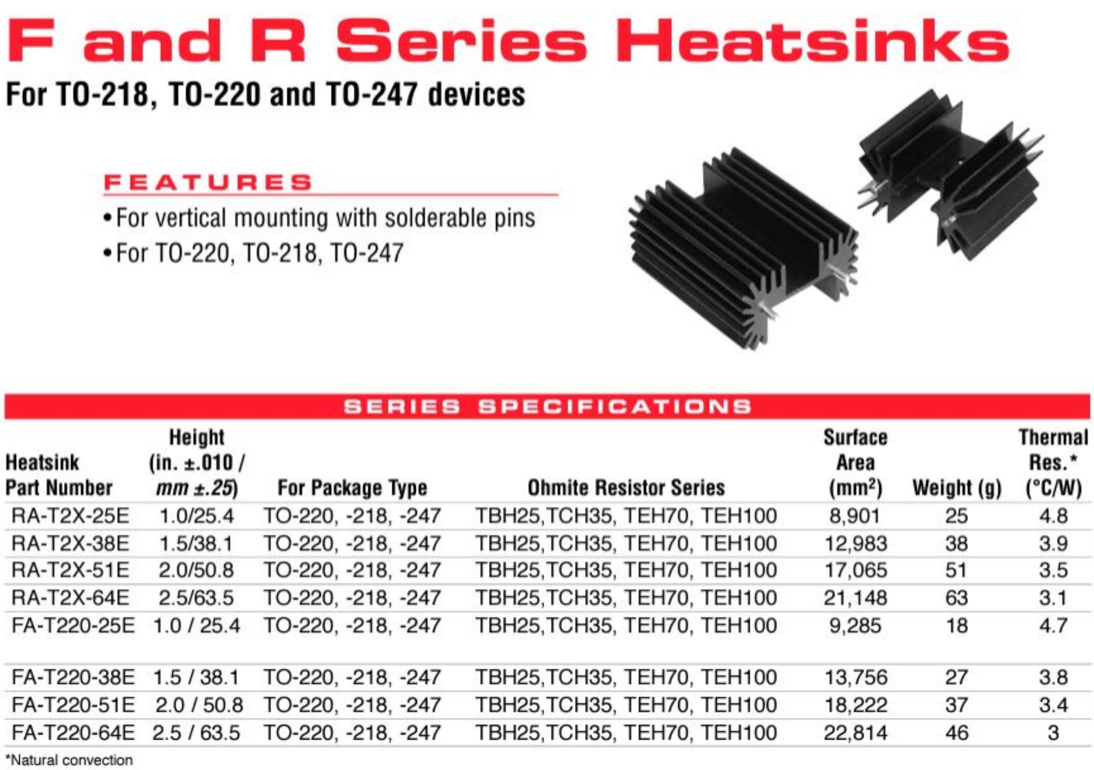
$$\text{Total Thermal Resistance} = 9.82°C/W$$

Divide the Total Thermal Resistance into the Designed Maximum Temperature to find the Maximum Power for the HSE-B20635-035H.

$$\text{HSE-B20635-035H}_{MaxPower} = \frac{MaxTemp_{Design}}{ThermalR_{Total}}$$

$$\text{HSE-B20635-035H}_{MaxPower} = \frac{110°C}{9.82°C/W}$$

✓ $\text{HSE-B20635-035H}_{MaxPower} = 11.2\ Watts$



F and R Series Heatsinks

For TO-218, TO-220 and TO-247 devices

FEATURES

- For vertical mounting with solderable pins
- For TO-220, TO-218, TO-247

SERIES SPECIFICATIONS

Heatsink Part Number	Height (in. ±.010 / *mm ±.25*)	For Package Type	Ohmite Resistor Series	Surface Area (mm²)	Weight (g)	Thermal Res.* (°C/W)
RA-T2X-25E	1.0/25.4	TO-220, -218, -247	TBH25,TCH35, TEH70, TEH100	8,901	25	4.8
RA-T2X-38E	1.5/38.1	TO-220, -218, -247	TBH25,TCH35, TEH70, TEH100	12,983	38	3.9
RA-T2X-51E	2.0/50.8	TO-220, -218, -247	TBH25,TCH35, TEH70, TEH100	17,065	51	3.5
RA-T2X-64E	2.5/63.5	TO-220, -218, -247	TBH25,TCH35, TEH70, TEH100	21,148	63	3.1
FA-T220-25E	1.0 / 25.4	TO-220, -218, -247	TBH25,TCH35, TEH70, TEH100	9,285	18	4.7
FA-T220-38E	1.5 / 38.1	TO-220, -218, -247	TBH25,TCH35, TEH70, TEH100	13,756	27	3.8
FA-T220-51E	2.0 / 50.8	TO-220, -218, -247	TBH25,TCH35, TEH70, TEH100	18,222	37	3.4
FA-T220-64E	2.5 / 63.5	TO-220, -218, -247	TBH25,TCH35, TEH70, TEH100	22,814	46	3

*Natural convection

Figure 5c.4: RA-T2X series Heat Sink data

5c.2.4 RA-T2X-25E

When using a heat sink, add the Thermal Resistance in series to find the Total Thermal Resistance between the device and the heat sink.

RA-T2X-25E Thermal Resistance = $4.8°C/W$

TIP41 Thermal Resistance Junction to Case= $1.67°C/W$

Total Thermal Resistance = $4.8°C/W + 1.67°C/W$

Total Thermal Resistance = $6.47°C/W$

Divide the Total Thermal Resistance into the Designed Maximum Temperature to find the Maximum Power for the RA-T2X-25E.

RA-T2X-25E$_{MaxPower}$ = $\frac{MaxTemp_{Design}}{ThermalR_{Total}}$

RA-T2X-25E$_{MaxPower}$ = $\frac{110°C}{6.47°C/W}$

✓ RA-T2X-25E$_{MaxPower}$ = $17\ Watts$

5c.2.5 RA-T2X-64E

When using a heat sink, add the Thermal Resistance in series to find the Total Thermal Resistance between the device and the heat sink.

RA-T2X-64E Thermal Resistance = $3.1°C/W$

TIP41 Thermal Resistance Junction to Case= $1.67°C/W$

Total Thermal Resistance = $3.1°C/W + 1.67°C/W$

Total Thermal Resistance = $4.77°C/W$

Divide the Total Thermal Resistance into the Designed Maximum Temperature to find the Maximum Power for the RA-T2X-64E.

RA-T2X-64E$_{MaxPower}$ = $\frac{MaxTemp_{Design}}{ThermalR_{Total}}$

RA-T2X-64E$_{MaxPower}$ = $\frac{110°C}{4.77°C/W}$

✓ RA-T2X-64E$_{MaxPower}$ = $23\ Watts$

Table 5c.1: Heat Sink Max Power Dissipation Calculations for the TIP41 T0220.

Heat Sink	Total Thermal Resistance	$Power_{Max} = \frac{110}{ThermalR_{Total}}$
TIP41 (No Heat Sink)	$57°C/W$	$1.930W$
HSE-B20254-035H	$14.6°C/W$	$7.534W$
HSE-B20635-035H	$9.82°C/W$	$11.202W$
RA-T2X-25E	$6.47°C/W$	$17.002W$
RA-T2X-64E	$4.77°C/W$	$23.061W$

Week 6

Differential Amplifiers

6.1 Objectives

Differential Amplifiers:

- Define Differential Amplification and Operating Modes.
- Design and design considerations for both a Differential Amplifier with an R_{Tail} and a Differential Amplifier with Constant Current Source.
- Analysis of Differential Amplifiers to include Calculating DC Currents and Voltages, Z_{IN}, Z_{OUT}, and Gains.
- Calculate the Common Mode Rejection Ratio
- Implementation of Differential Amplifier Gain Control.

6.2 Differential Amplifier Introduction

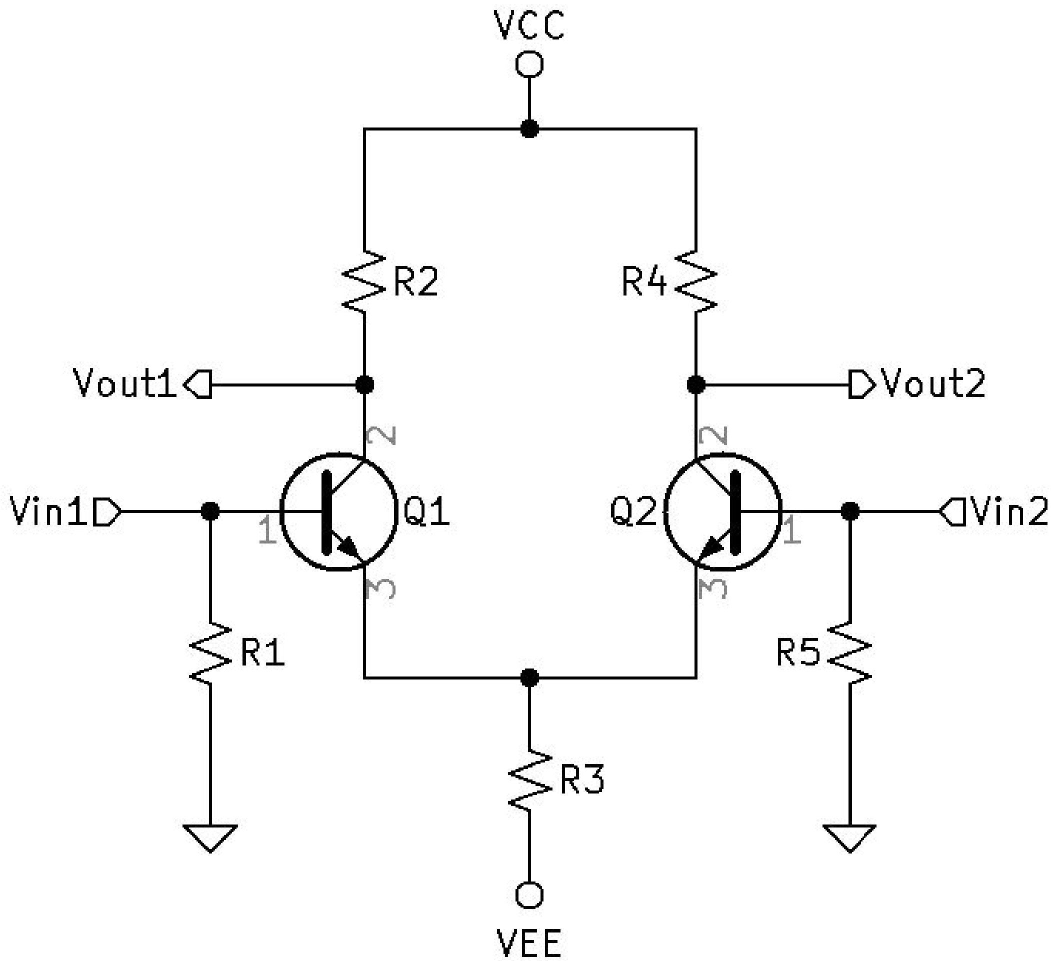


Figure 6.1: Differential Amplifier Introduction

A **Differential Amplifier** is an amplifier that produces outputs that are a function of the difference between the two inputs.

Refer to Figure 6.1.

- ✓ Transistors Q1 and Q2 are Beta matched.
- ✓ The Collector Resistors are matched (R2 = R4).
- ✓ The Base Resistors are matched (R1 = R5).
- ✓ The tail current, IR3 is equal to 2xIE.

6.3 Differential Amplifier Operational Modes

6.3.1 Single-Ended Differential Input

The Differential Amplifier is operated with one input grounded while signal voltage is applied to the other input.

Single-Ended Differential Input *Inverting Amplification*

The output waveform is 180° out of phase with the input.

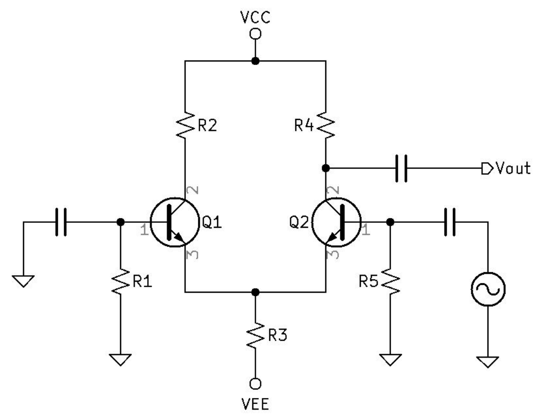


Figure 6.2: Differential Amplifier Single-Ended Differential Input *Inverting Amplifier*

Single-Ended Differential Input *Non-Inverting Amplification*

The output waveform is in phase with the input.

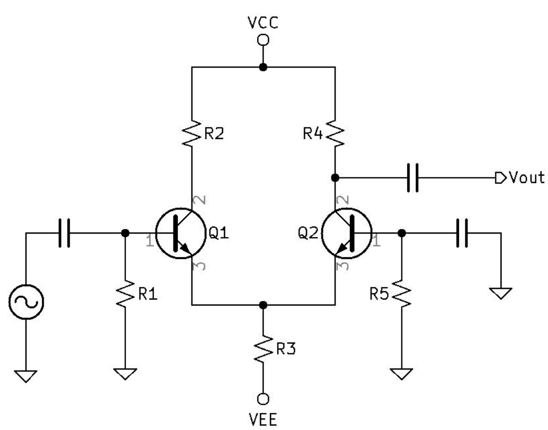


Figure 6.3: Differential Amplifier Single-Ended Differential Input *Non-Inverting Amplifier*

6.3.2 Double-Ended Differential Inputs

The two inputs are 180° out of phase with each other and the gain doubled, $2 \times \Delta V$.

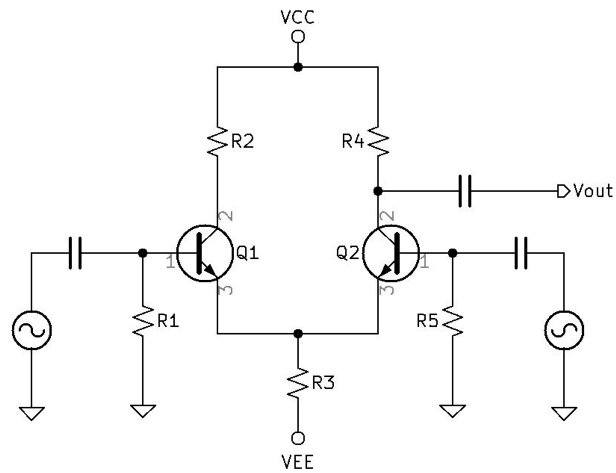


Figure 6.4: Differential Amplifier Double-Ended Differential Input

Double-Ended Common-Mode Inputs

The two inputs are in phase with each other and the gain is approximately zero, $\Delta V \approx 0$.

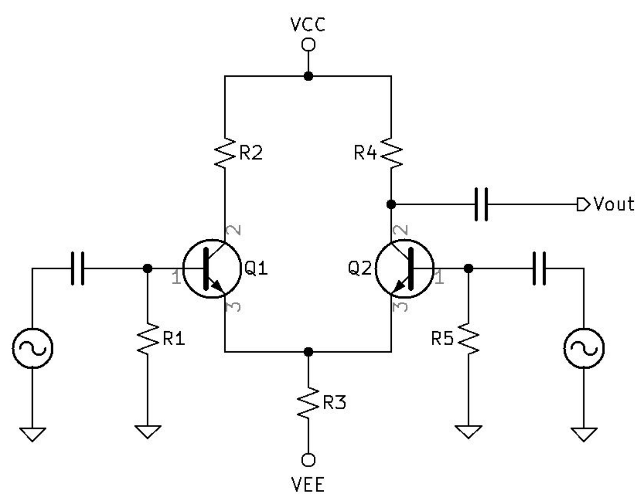


Figure 6.5: Differential Amplifier Double-Ended Common-Mode Input

Differential Amplification XLR Example

The beauty of the Differential Amplifier is that can achieve the benefits of both Differential and Common-Mode amplification within the same application and at the same time. Imagine a mono guitar signal is sent to a Diff-Amp and Right (out of phase) and Left (in phase) signals are produced and sent down a long XLR cable. On the way the XLR cable will pick up lots of noise, that noise will be in phase on both the Left and Right channels. At the other end of the cable, a second Diff-Amp is ready to provide Diff Gain $(2 \times \Delta V)$ to the original guitar signal and Common-Mode gain of approximately zero to the in-phase noise that was picked up along the way.

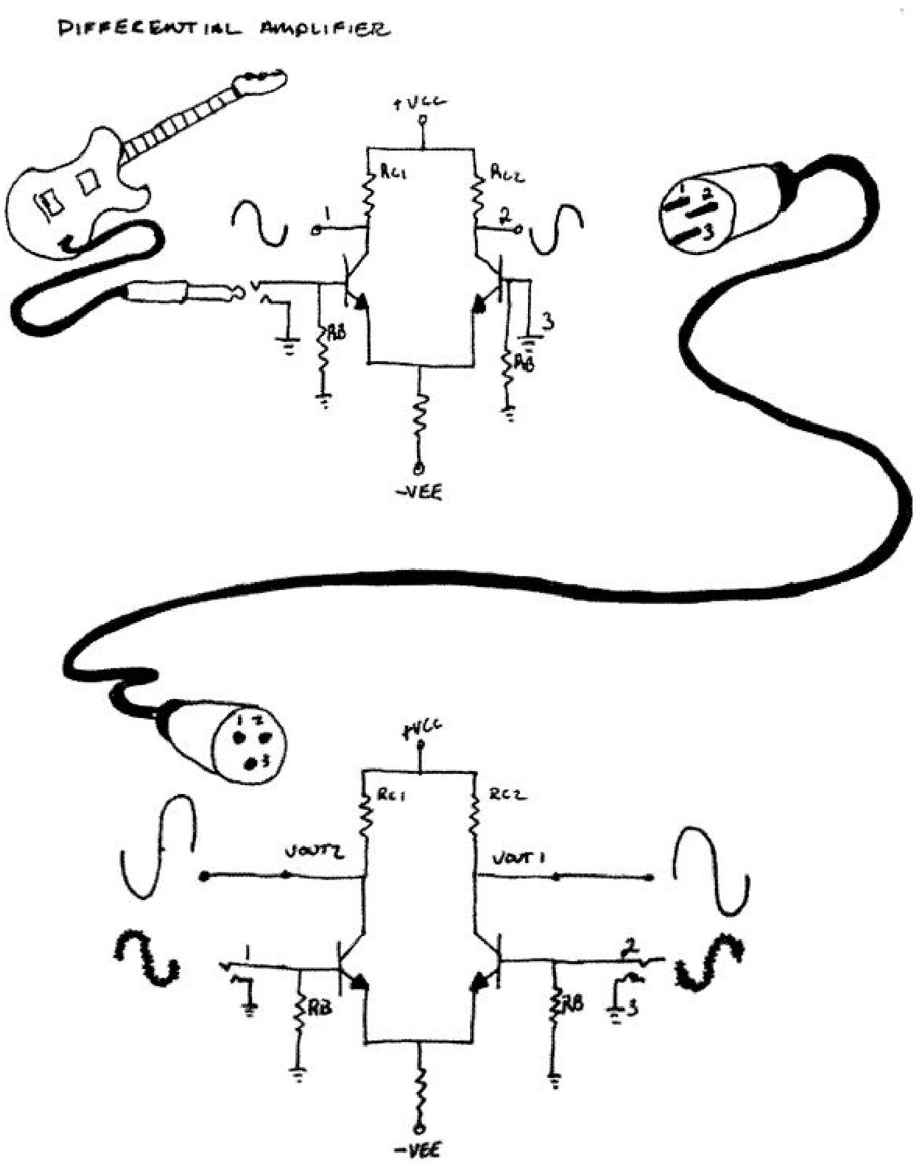


Figure 6.6: Differential Amplifiers: Concurrent $2 \times \Delta V$ with Common-Mode noise rejection.

6.4 Differential Amplifier R-Tail Practical Design

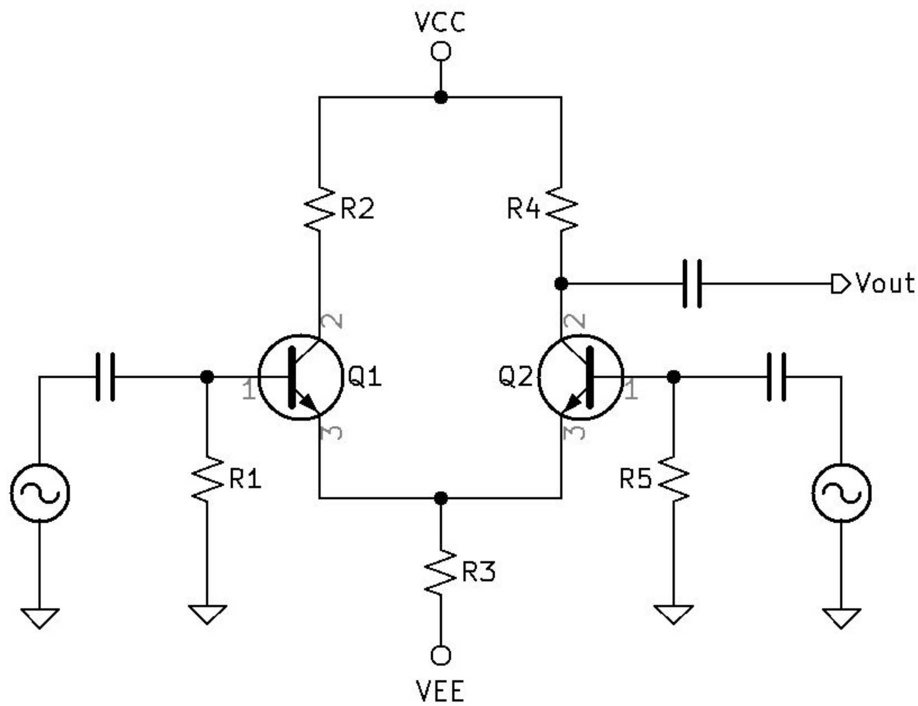


Figure 6.7: Differential Amplifier with Tail Resistor

DC Biasing

Maximum Desired Peak Output Voltage

Determining the Maximum Peak Output Voltage needed for the amplifier.

$$Vout_{MaxP} = ?$$

Determine VCC and VEE based on Maximum Desired Peak Output Voltage

If the amplifier is biased correctly, using the following procedure, 45% of the total DC voltage (VCC+VEE) will be approximately equal to the Maximum Peak Output Voltage.

$$Vout_{MaxP} \approx 45\%(VCC + VEE)$$

Solve for $VCC + VEE$.

$$VCC + VEE \approx \frac{Vout_{MaxP}}{45\%}$$

Making VCC equal to VEE, we can solve for VCC in terms of our desired Maximum Peak Output Voltage.

$$VCC + VCC \approx \frac{Vout_{MaxP}}{45\%}$$

$$2VCC \approx \frac{Vout_{MaxP}}{45\%}$$

$$VCC \approx \frac{Vout_{MaxP}}{2 \times 0.45}$$

Headroom, adjusting VCC and VEE to prevent amplifier distortion

To prevent undesired distortion caused by the amplifier operating near Vout Max Peak, we need to add headroom to the biasing circuit. For design purposes, 10% of additional should be adequate. This means multiplying the calculated VCC by 110%

$$VCC = \frac{Vout_{MaxP}}{2 \times 0.45} \times 1.1$$

VEE is equal to VCC but negative in polarity.

$$VEE = -VCC$$

Voltage Optimization 45, 45, 10

The total bias voltage is equal to $VCC + VEE$.

$$VDC_{Total} = VEE + VCC$$

Make VRC equal to 45% of the total bias voltage.

$$V_{RC} = 0.45 \times VDC_{Total}$$

Make VCE equal to 45% of the total bias voltage.

$$V_{CE} = 0.45 \times VDC_{Total}$$

Make the Tail Voltage equal to 10% of the total bias voltage.

$$V_{RTail} = 0.1 \times VDC_{Total}$$

The Collector Resistor

The Ideal output impedance of any amplifier is low. Because Zout is equal to RC, we want RC to be as small as possible. We know VRC is equal to VCE. Using the power rating of the transistor, we can determine a suitable RC value.

Look up the Maximum Power Rating for the transistor.

Set the transistor power to safely operate at half of its max power.

$P_Q = \frac{P_{Q1Max}}{2}$

Because VCE and VRC are equal and they have IC in common, they will also dissipate the same power.

$P_{RC} = P_Q$

Calculate RC in terms of Voltage and Power.

$P = \frac{V^2}{R}$

$P_{RC} = \frac{(V_{RC})^2}{RC}$

$RC = \frac{(V_{RC})^2}{P_{RC}}$

Solve for IC in terms of Power and Resistance.

$P = I^2 \times R$

$P_{RC} = (I_{RC})^2 \times RC$

$(I_{RC})^2 = \frac{P_{RC}}{I_{RC}}$

$I_{RC} = \sqrt{\frac{P_{RC}}{I_{RC}}}$

Referencing Figure 6.8. R2 and R4 are equal to RC.

$R2 = R4 = RC$

The Base Resistor

Using Kirchhoff's Voltage Law, calculate VRB.

$$-V_{RTail} - V_{BE} - V_{RB} + VEE = 0$$

$$V_{RB} = VEE - V_{BE} - V_{RTail}$$

Solve for IB.

$$IB = \frac{IC}{\beta}$$

Note: Differential Amplifier transistors, ideally, are beta-matched. Also, The Differential Amplifier circuit is Beta dependent.

Solve for RB using Ohm's Law.

$$RB = \frac{V_{RB}}{I_{RB}}$$

Reference Figure 6.8. $R1 = R5 = RB$

The Tail Resistor

Using Kirchhoff, the Tail Resistor Current is equal to $2 \times IE$.

Find IE

$$IE = IB \times (\beta + 1)$$

Find the Tail Resistor Current

$$I_{Tail} = 2 \times IE$$

The dynamic emitter resistance (r'e)

$$r'e \approx \frac{0.026}{IE}$$

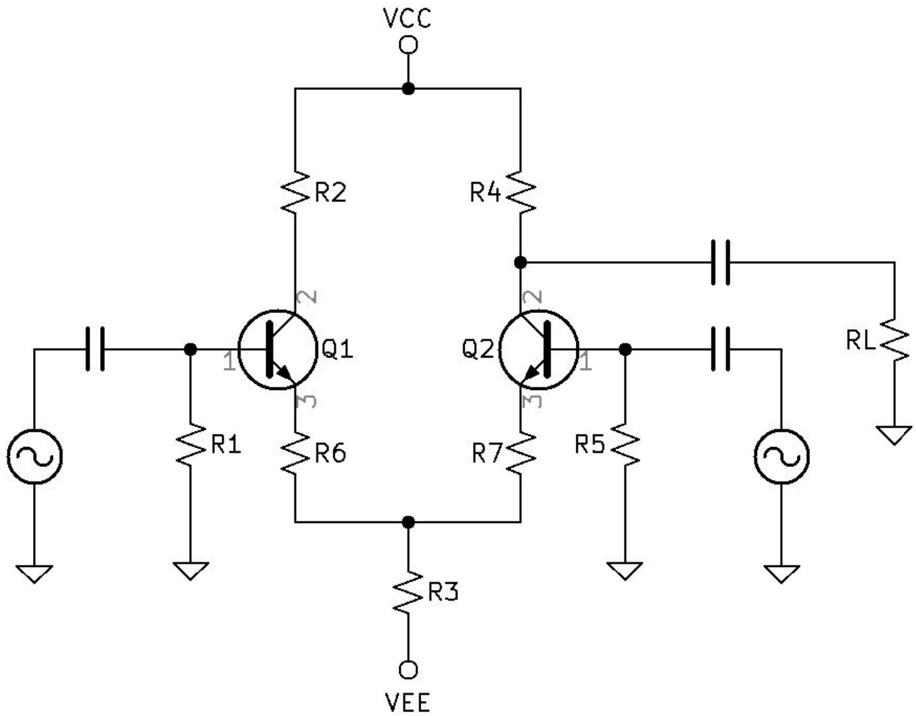


Figure 6.8: Differential Amplifier with Swamp Resistors and Load Resistance

Swamping Resistors

Adding the Swamping Resistors R6 and R7 will improve stability and Zin of the Differential Amplifier, while reducing the dependency on r'e.

$$R_{Swamp} = 10 \times r'e$$

$$r'e = \frac{0.026}{IE}$$

$$R6 = R7 = R_{Swamp}$$

Load Resistance

To Maximize Vout Peak, ensure that RL is ten times larger than Zout (R4) of the Differential Amplifier. Understand that RL may also represent the Z_{In} of the next stage, perhaps a Push-Pull Amplifier or an additional voltage gain stage if needed.

$$RL = 10 \times Zout_{DiffAmp}$$

$$RL = 10 \times R4$$

$$RL = Zin_{Stage2}$$

$$Zin_{Stage2} \geq 10 \times R4$$

6.5 Differential Amplifier R-Tail Calculations

Input Impedance

Refer to Figure 6.8. Z_{In} is equal to the impedance that the Generator sees.

$$Z_{In} = [(((\frac{R_{Gen}//RB}{\beta+1} + r'e + R_{Swamp})//R_{Tail}) + R_{Swamp} + r'e) \times (\beta + 1)]//RB$$

Output Impedance

The Output Impedance Z_{Out} is equal to the Impedance that the Load sees.

$$Z_{Out} = RC$$

Single-Ended Inverting Voltage Gain

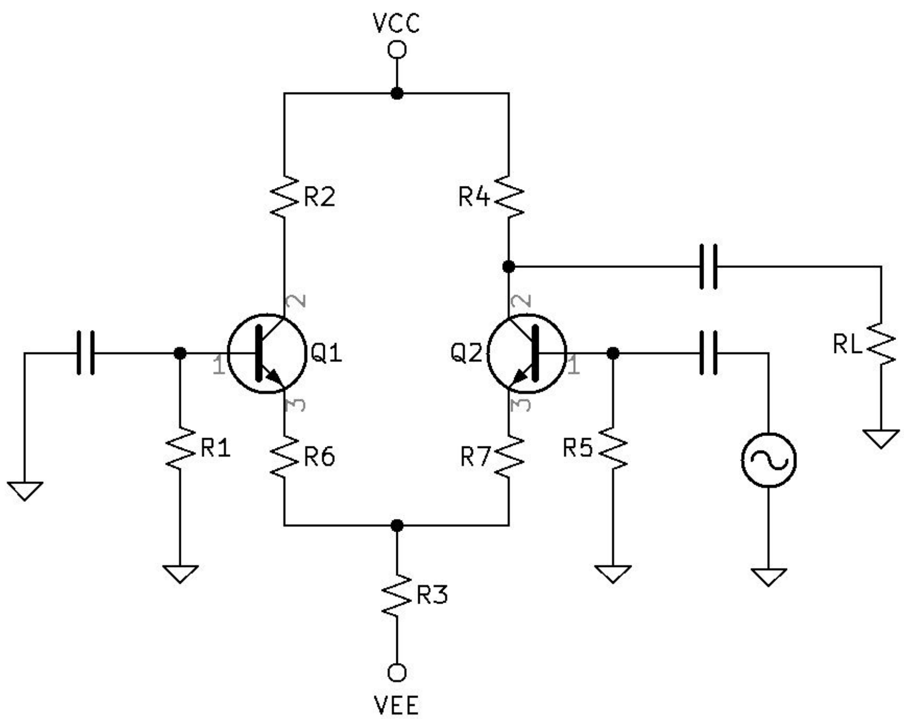


Figure 6.9: Single-Ended Inverting Differential Amplifier

$$\Delta V_{CE} = \frac{Vout}{Vin}$$

$$\Delta V_{CE} = \frac{IC(RL//RC)}{IB(((r'e+R_{Swamp})//R_{Tail})+r'e+R_{Swamp})\times(\beta+1)}$$

$$\Delta V_{CE} = \alpha\frac{RL//RC}{((r'e+R_{Swamp})//R_{Tail})+r'e+R_{Swamp}}$$

Single-Ended Non-Inverting Voltage Gain

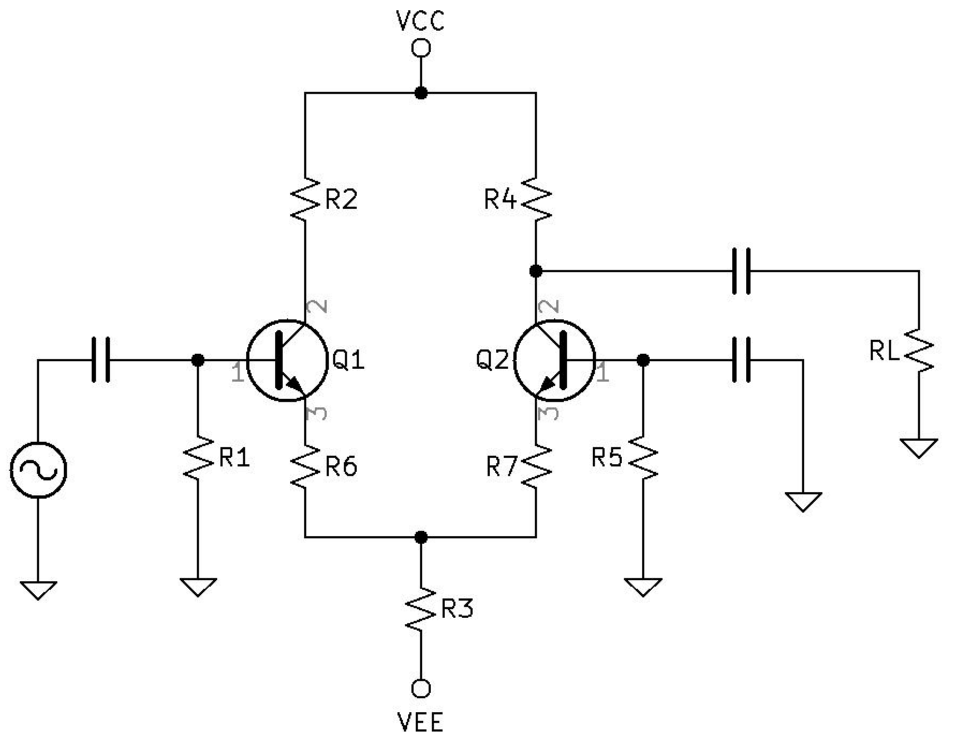


Figure 6.10: Single-Ended Non-Inverting Differential Amplifier

$\Delta V_{CC} = \frac{Vout}{Vin} \approx 0.5$

$$\Delta V_{CC} = \frac{IE((r'e+R_{Swamp})//R_{Tail}}{IB(((r'e+R_{Swamp})//R_{Tail})+R_{Swamp}+r'e)\times(\beta+1)}$$

$$\Delta V_{CC} = \frac{IE((r'e+R_{Swamp})//R_{Tail}}{IE(((r'e+R_{Swamp})//R_{Tail})+R_{Swamp}+r'e)}$$

$$\Delta V_{CC} = \frac{(r'e+R_{Swamp})//R_{Tail}}{((r'e+R_{Swamp})//R_{Tail})+R_{Swamp}+r'e}$$

$\Delta V_{CB} = \frac{Vout}{Vin} \approx 2 \times \Delta V_{CE}$

$$\Delta V_{CB} = \frac{IC(RL//RC)}{IE(R_{Swamp}+r'e)}$$

$$\Delta V_{CB} = \alpha\frac{RL//RC}{R_{Swamp}+r'e}$$

$\Delta V_{CCCB} = \Delta V_{CC} \times \Delta V_{CB}$

Double-Ended Gain

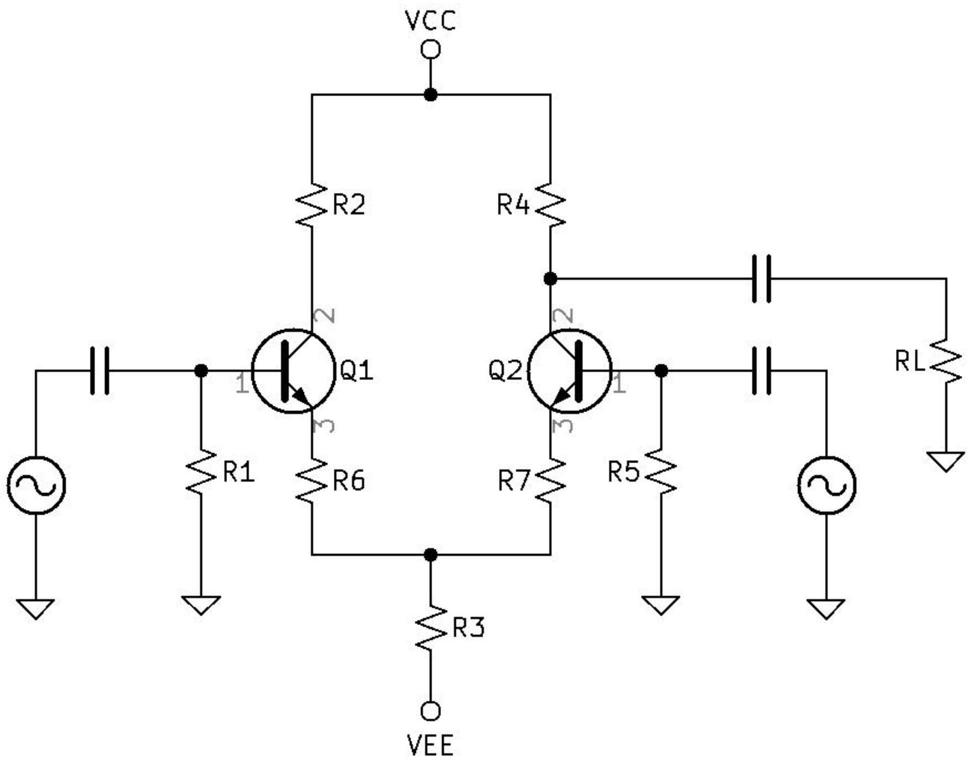


Figure 6.11: Double-Ended Common-Mode Differential Amplifier

Common-Mode

In Common-Mode, the two input signals are in phase will essentially cancel each-other at the output.

$$\Delta V_{CM} \approx 0$$

$$\Delta V_{CM} = \Delta V_{CCCB} - |\Delta V_{CE}|$$

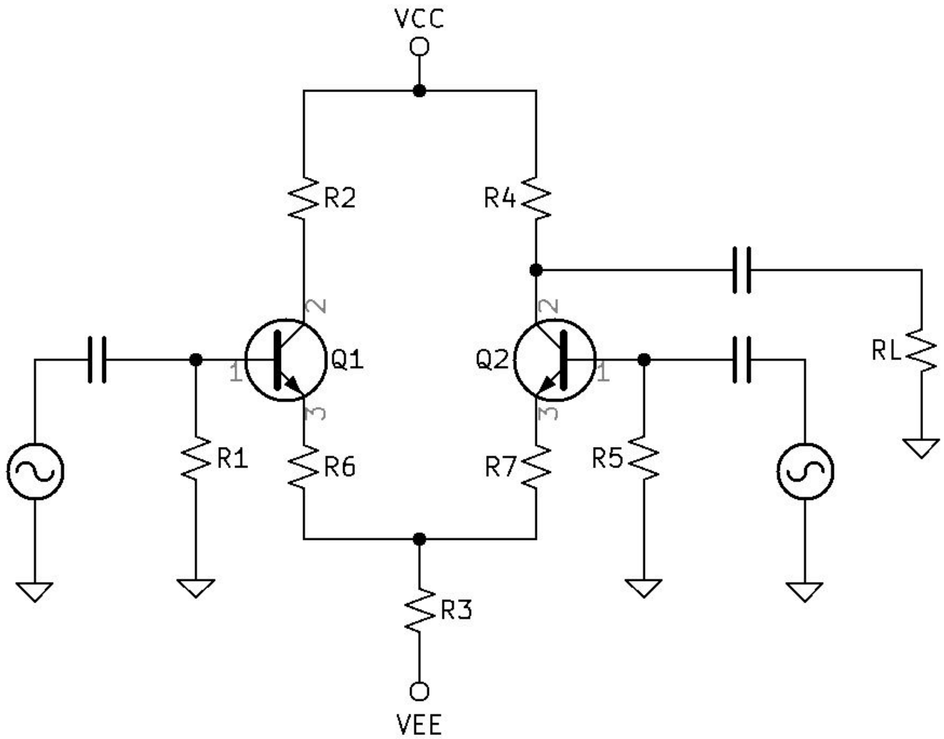


Figure 6.12: Double-Ended Differential-Mode Differential Amplifier

Differential-Mode

In Differential-Mode, the two input signals are out of phase will essentially double the signal voltage gain at the output.

$\Delta V_{DM} \approx 2 \times \Delta V_{CE}$

$\Delta V_{DM} = \Delta V_{CCCB} + |\Delta V_{CE}|$

Common Mode Rejection Ratio

The Common Mode Rejection Ration is the ratio of Common-Mode signal to Differential-Mode signal.

$CMMR = \frac{\Delta V_{CM}}{\Delta V_{DM}}$

Common Mode Rejection Ratio in decibels

$CMMR_{dB} = 20Log\frac{\Delta V_{CM}}{\Delta V_{DM}}$

6.6 Differential Amplifier Constant Current Source Practical Design

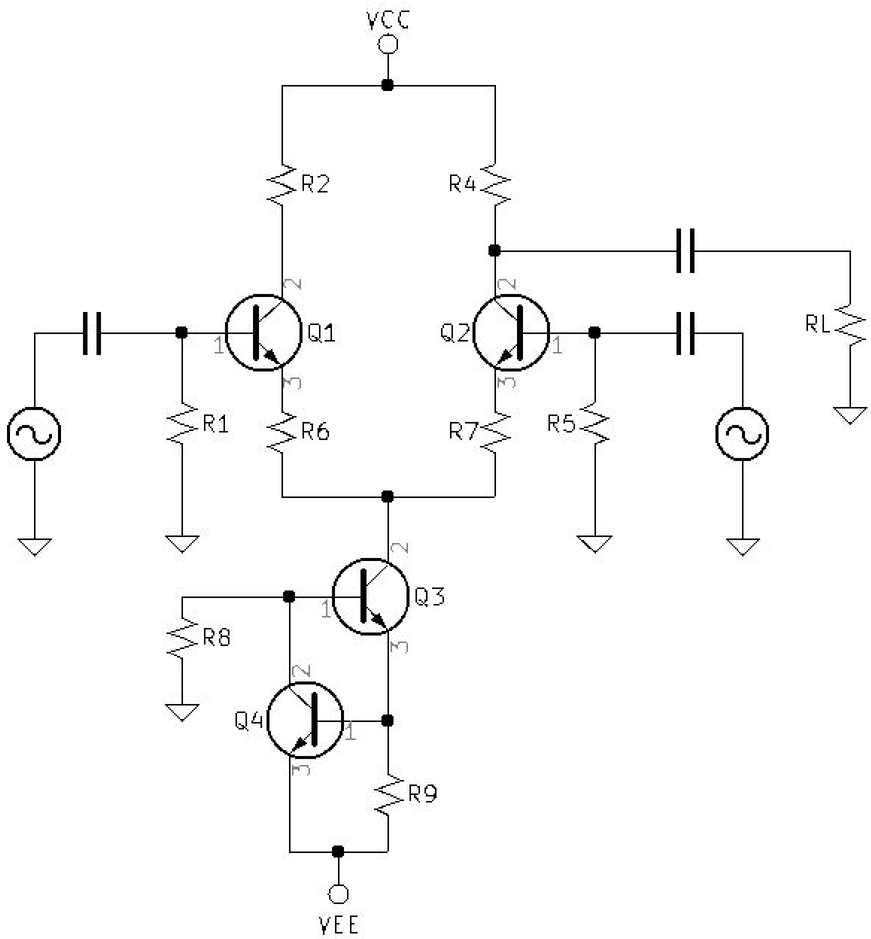


Figure 6.13: Differential Amplifier with a Constant Current Source

Reference Figure 6.13. Notice that R3 the Tail Resistor from the previous section has been replaced with the Constant Current Source which is comprised of Q3, Q4, R8, and R9.

VCC, VEE, Base Resistors, Collector Resistors, and Swamping Resistors are all calculated in the previous section.

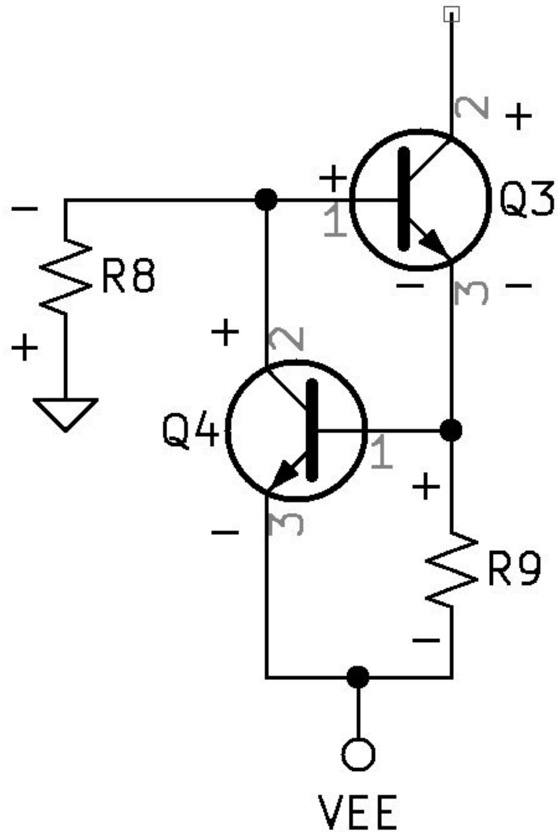


Figure 6.14: Constant Current Source

DC Biasing:

Q3 and Q4 Beta Match

Beta Q3 and Q4 are equal or nearly equal.

$$\beta_{Q3} \approx \beta_{Q4}$$

Designing in Terms of Tail Current

The collector current of Q3 is equal to the desired tail current.

$$IC_{Q3} = I_{Tail} = 2 \times IE$$

The base current of Q3 is equal to the desired tail current divided by beta.

$$IB_{Q3} = \frac{I_{Tail}}{\beta}$$

The emitter current of Q3 is equal to the desired tail current (IC) plus the base current of Q3.

$$IE_{Q3} = IC_{Q3} + IB_{Q3}$$

$$IE_{Q3} = I_{Tail} + \frac{I_{Tail}}{\beta}$$

Notice the emitter current of Q3 equation can also be written as being is equal to I_{R9} plus IB_{Q4}

$$IE_{Q3} = I_{R9} + IB_{Q4}$$

R9

Make the current of R9 equal to the tail current.

$$I_{R9} = I_{Tail}$$

Use Kirchhoff's Voltage Law to find the R9 Voltage.

$$V_{R9} = VBE_{Q4}$$

$$V_{R9} = 0.7V$$

Use Ohm's Law to find the R9.

$$R9 = \frac{V_{R9}}{I_{R9}}$$

$$R9 = \frac{0.7V}{I_{tail}}$$

R8

Find IB Q4. If the emitter current of Q3 is equal to the tail current plus the tail current divided by beta and IR9 is now equal to the tail current, then the base current of Q4 must be equal to the tail current divided by beta. Which is also equal to IB_{Q3}.

$$IE_{Q3} = IC_{Q3} + IB_{Q3}$$

$$IE_{Q3} = I_{Tail} + \frac{I_{Tail}}{\beta}$$

$$IE_{Q3} = I_{R9} + \frac{I_{Tail}}{\beta}$$

$$IB_{Q4} = \frac{I_{Tail}}{\beta}$$

$$IB_{Q4} = IB_{Q3}$$

Find IC Q4. Again, assuming that Q3 and Q4 are beta matched, then the betas cancel and IC Q4 is equal to the Tail Current.

$$IC_{Q4} = IB_{Q4} \times \beta$$

$$IC_{Q4} = \frac{I_{Tail}}{\beta} \times \beta$$

$$IC_{Q4} = I_{Tail}$$

Find R8's current. IR8 is equal to the base current of Q3 plus the collector current of Q4.

$$I_{R8} = IB_{Q3} + IC_{Q4}$$

$$I_{R8} = \frac{I_{tail}}{\beta} + I_{tail}$$

Factor out the I_{tail}:

$$I_{R8} = I_{tail}(\frac{1}{\beta} + 1)$$

$$I_{R8} \approx I_{tail}$$

Use Kirchhoff's Voltage Law to calculate the R8 Voltage.

$$V_{R8} + VBE_{Q3} + VBE_{Q4} - VEE = 0$$

$$V_{R8} = VEE - VBE_{Q3} - VBE_{Q4}$$

$$V_{R8} = VEE - 1.4V$$

Find R8 using Ohm's Law.

$$R8 = \frac{V_{R8}}{I_{R8}}$$

$$R8 = \frac{VEE-1.4V}{I_{tail}(\frac{1}{\beta}+1)} \approx \frac{VEE-1.4V}{I_{tail}}$$

6.7 Differential Amplifier Constant Current Source Calculations

Input Impedance

Refer to Figure 6.13. Z_{In} is equal to the impedance that the Generator sees. Notice the collector resistance of Q3 will appear as open or infinitely large.

$$Z_{In} = ((\frac{R_{Gen}//RB}{\beta+1} + r'e + R_{Swamp} + R_{Swamp} + r'e) \times (\beta + 1))//RB$$

Output Impedance

The Output Impedance Z_{Out} is equal to the Impedance that the Load sees.

$$Z_{Out} = RC$$

Load

Make the load (RL) a minimum of ten times larger than Z_{Out}.

Single-Ended Gains:

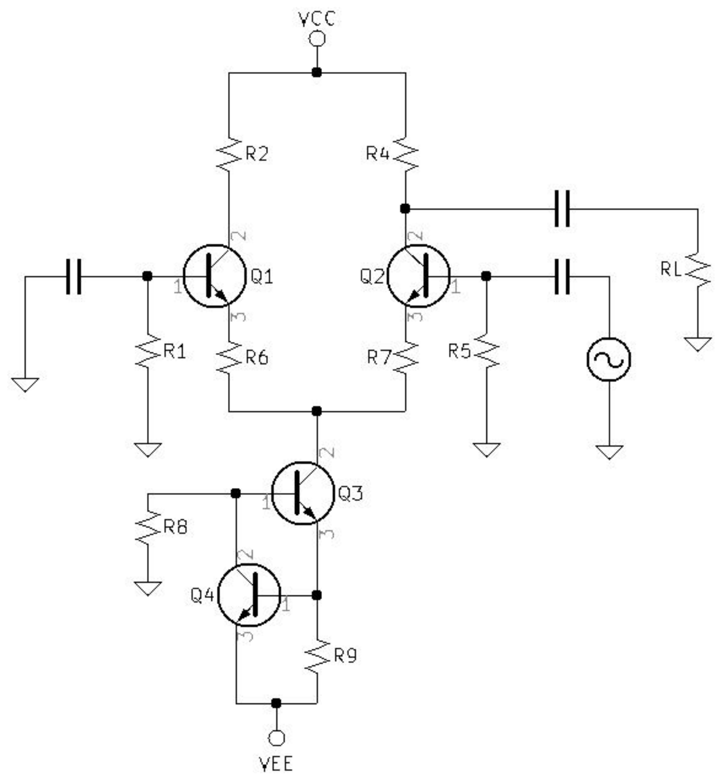


Figure 6.15: Single-Ended Inverting Differential Amplifier with Constant Current Source

Single-Ended Inverting Voltage Gain

See Figure 6.15.

$$\Delta V_{CE} = \frac{Vout}{Vin}$$

$$\Delta V_{CE} = \frac{IC(RL//RC)}{IB((r'e+R_{Swamp})+r'e+R_{Swamp})\times(\beta+1)}$$

$$\Delta V_{CE} = \alpha\frac{RL//RC}{2r'e+2R_{Swamp}}$$

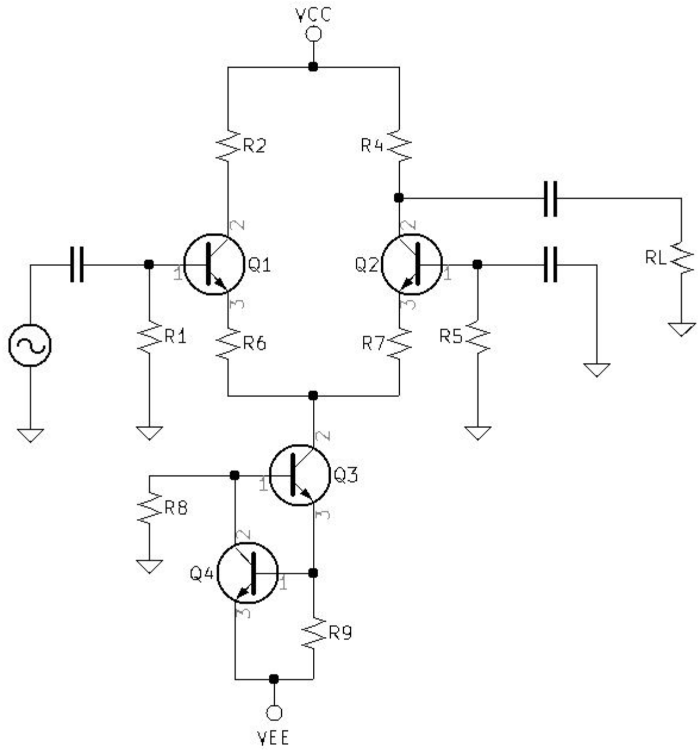


Figure 6.16: Single-Ended Non-Inverting Differential Amplifier with Constant Current Source

Single-Ended Non-Inverting Voltage Gain

See Figure 6.16.

$$\Delta V_{CC} = \frac{Vout}{Vin}$$

$$\Delta V_{CC} = \frac{IE(r'e+R_{Swamp})}{IB(r'e+R_{Swamp}+R_{Swamp}+r'e)\times(\beta+1)}$$

$$\Delta V_{CC} = \frac{IE((r'e+R_{Swamp})}{IE(r'e+R_{Swamp}+R_{Swamp}+r'e)}$$

$$\Delta V_{CC} = \frac{r'e+R_{Swamp}}{2r'e+2R_{Swamp}}$$

$$\Delta V_{CC} = \frac{r'e+R_{Swamp}}{2(r'e+R_{Swamp})}$$

$$\Delta V_{CC} = \frac{1}{2} \times \frac{r'e+R_{Swamp}}{r'e+R_{Swamp}}$$

$$\Delta V_{CC} = \frac{1}{2} \times 1$$

$$\boxed{\Delta V_{CC} = \frac{1}{2} = 0.5}$$

$$\Delta V_{CB} = \frac{Vout}{Vin} \approx 2 \times \Delta V_{CE}$$

$$\Delta V_{CB} = \frac{IC(RL//RC)}{IE(R_{Swamp}+r'e)}$$

$$\boxed{\Delta V_{CB} = \alpha \frac{RL//RC}{R_{Swamp}+r'e}}$$

$$\Delta V_{CCCB} = \Delta V_{CC} \times \Delta V_{CB}$$

$$\Delta V_{CCCB} = 0.5 \times \Delta V_{CB}$$

Double-Ended Gains:

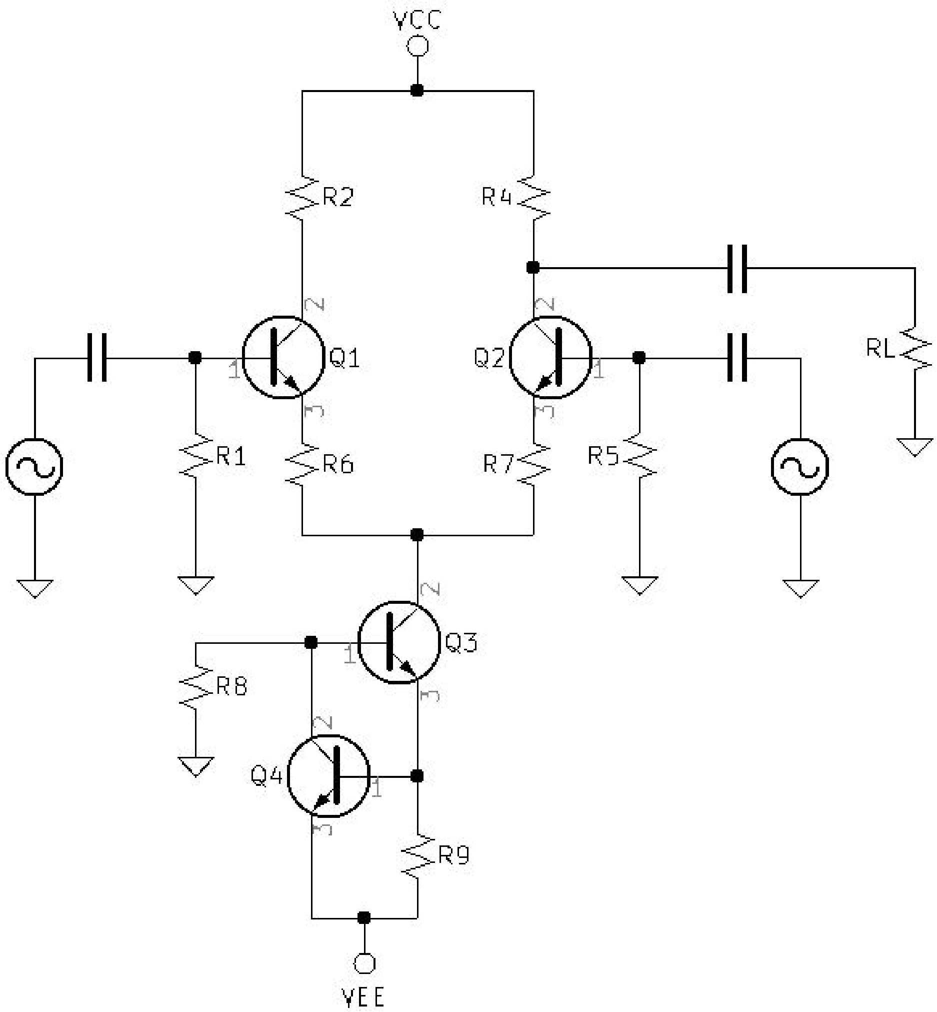


Figure 6.17: Double-Ended Common-Mode Differential Amplifier with Constant Current Source

Common-Mode

In Common-Mode, the two input signals are in phase and will essentially cancel each other at the output.

$$\Delta V_{CM} \approx 0$$

$$\Delta V_{CM} = \Delta V_{CCCB} - |\Delta V_{CE}|$$

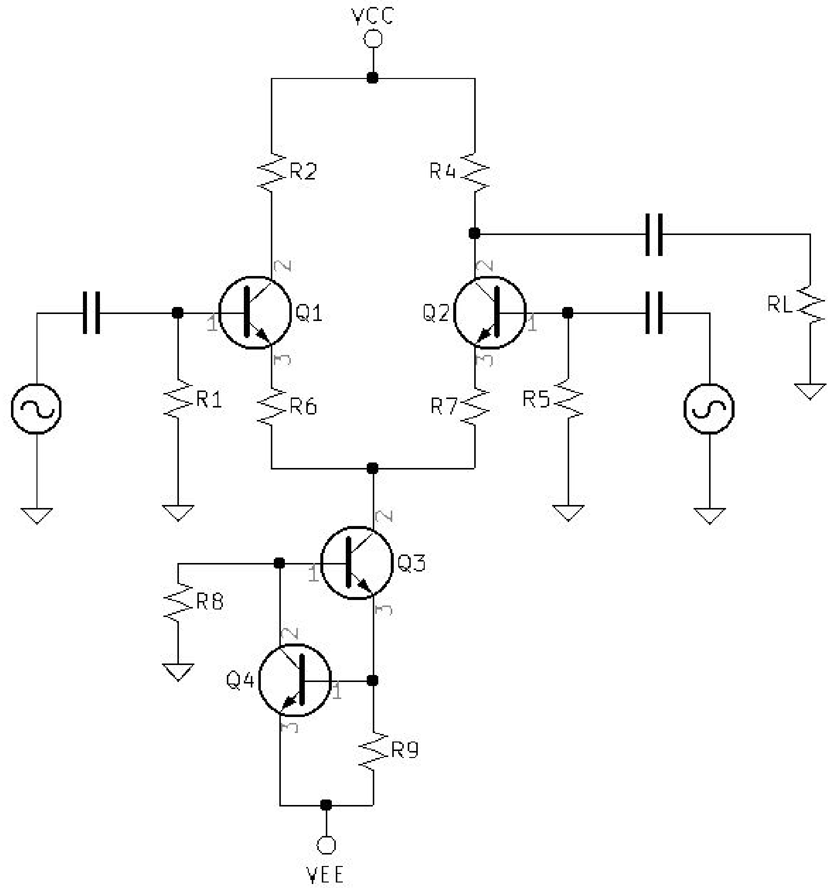


Figure 6.18: Double-Ended Differential-Mode Differential Amplifier with Constant Current Source

Differential-Mode

In Differential-Mode, the two input signals are out of phase will essentially double the signal voltage gain at the output.

$$\Delta V_{DM} \approx 2 \times \Delta V_{CE}$$

$$\Delta V_{DM} = \Delta V_{CCCB} + |\Delta V_{CE}|$$

Common Mode Rejection Ratio

The Common Mode Rejection Ratio is the ratio of Common-Mode signal to Differential-Mode signal.

$$CMRR = \frac{\Delta V_{CM}}{\Delta V_{DM}} \qquad CMRR_{dB} = 20Log\frac{\Delta V_{CM}}{\Delta V_{DM}}$$

6.8 Differential Amplifier Constant Current Source Practical Design

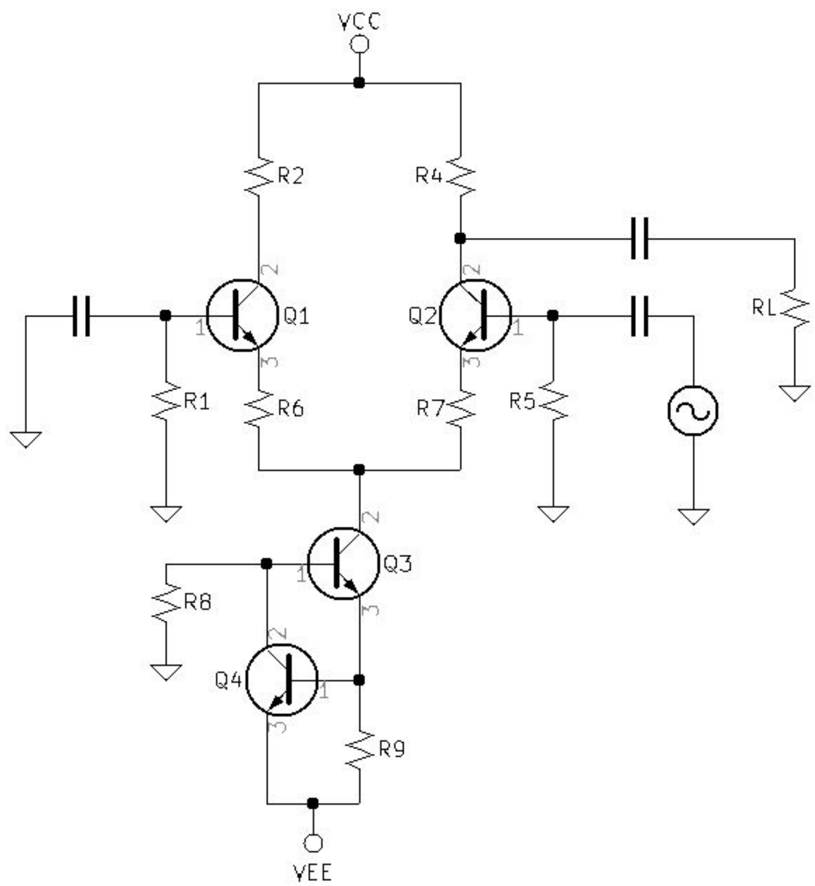


Figure 6.19: Differential Amplifier with a Constant Current Source and Open Loop Gain

Just like Operational Amplifiers, we can control the gain of the Differential Amplifier using negative feedback. This is accomplished by feeding back a fraction of the output signal back to the negative input.

6.8.1 Inverting Diff Amp with Gain Control

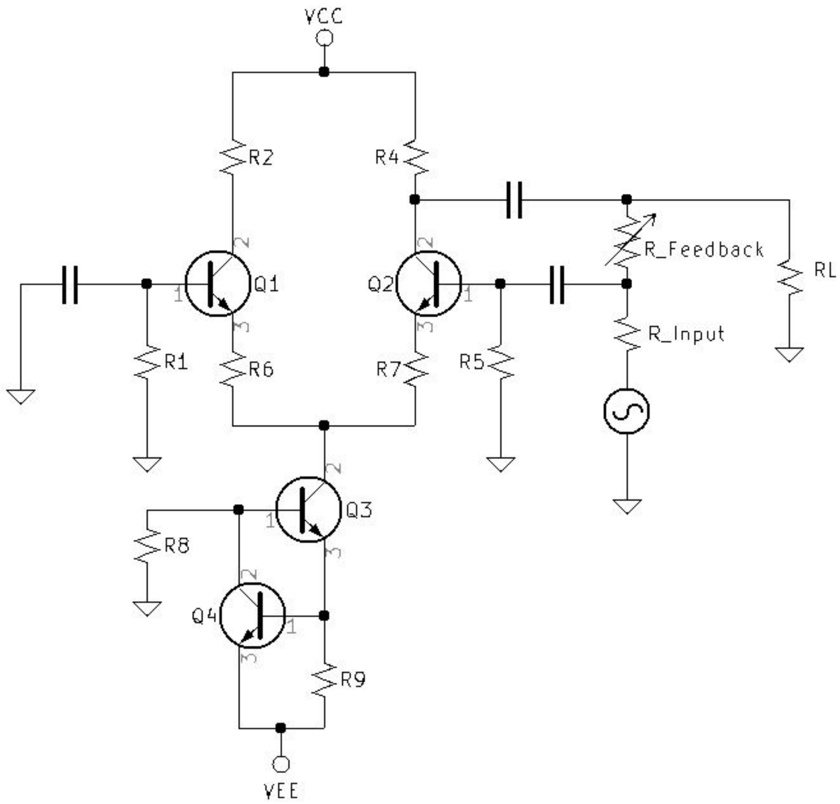


Figure 6.20: Inverting Differential Amplifier with Gain Control

See Figure 6.20. The Inverting Gain will be approximately equal to the feedback resistance divided by the input resistance.

$\Delta V \approx -\frac{R_{Feedback}}{R_{Input}}$

6.8.2 Non-Inverting Diff Amp with Gain Control

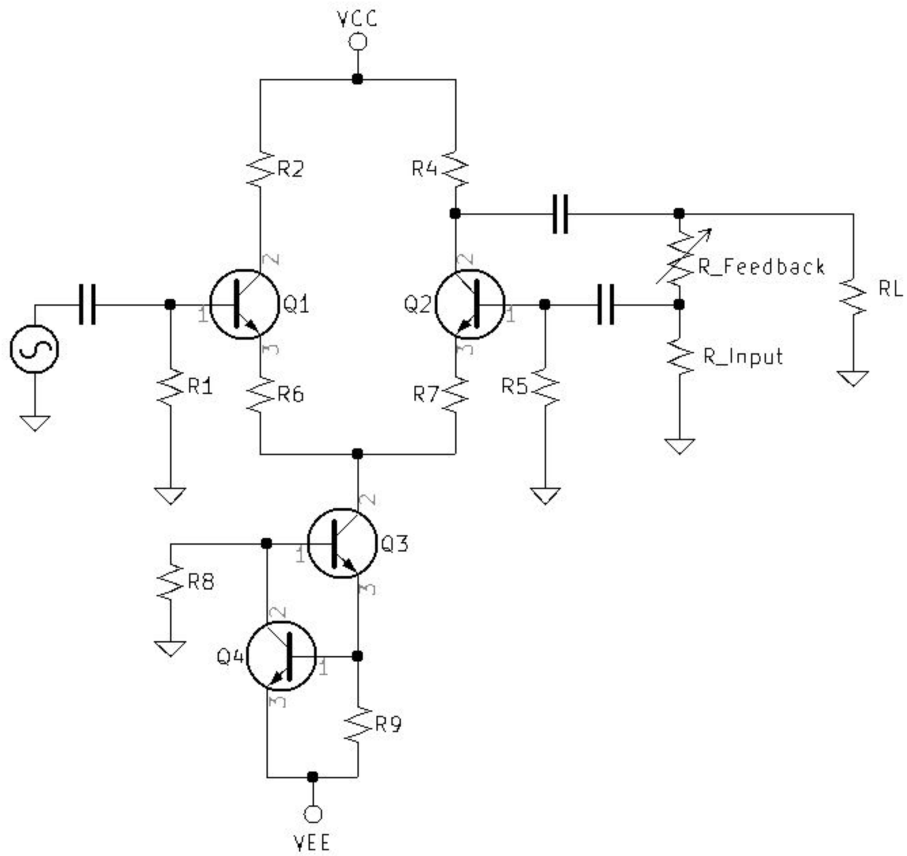


Figure 6.21: Non-Inverting Differential Amplifier with Gain Control

See Figure 6.21. The Non-Inverting Gain will be approximately equal to the feedback resistance divided by the input resistance plus one.

$\Delta V \approx \frac{R_{Feedback}}{R_{Input}} + 1$

6.8.3 Non-Inverting Diff Amp with Gain Control and Push-Pull Amplifier

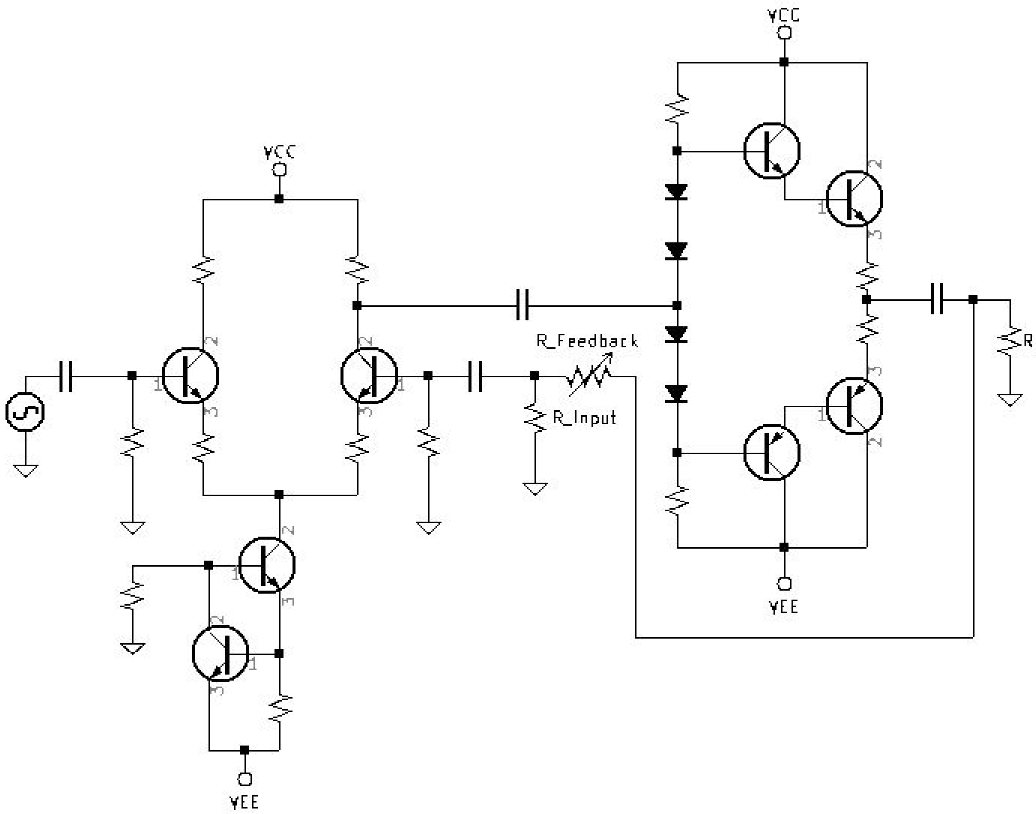


Figure 6.22: Non-Inverting Differential Amplifier with Gain Control and Push-Pull Amplifier

Figure 6.22 is an example of a Differential Amplifier using a Push-Pull Amplifier for current gain. For increased signal stability and voltage gain control, the negative feedback comes from the load.

Week 7

Operational Amplifiers

7.1 Objectives

Operational Amplifiers:

- Define the Ideal Characteristics Operational Amplifiers.
- Define the Practical Characteristics of the LM741 and TL071
- Define the Ideal Characteristics Operational Amplifiers.
- Identify the basic internal arrangement of an Operational Amplifier.
- Define Operational Amplifier Modes.
- Review the Rules and Steps for calculating Operational Amplifiers
- Identify and locate important specifications using the datasheet.
- Calculate Gains, Frequency Response, and Input & Output Impedances.
- Review Operational Amplifier Comparator Circuits.
- Explore Operational Amplifier Voltage Bounding Circuits.

7.2 Operational Amplifier General Information

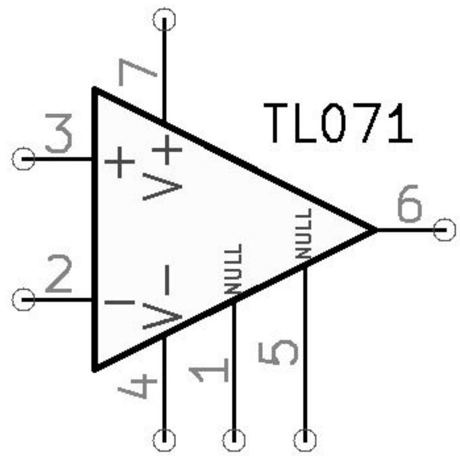


Figure 7.1: Operational Amplifier Introduction

7.2.1 The Ideal Characteristics of the Operational Amplifier

The Ideal Operational Amplifier will have the following characteristics:

- Infinite Gain ($\Delta V = \infty$)
- Infinite Bandwidth ($Frequency = 0\ to\ \infty hz$)
- Infinite Input Impedance ($Z_{in} = \infty\Omega$)
- Zero Output Impedance ($Z_{out} = 0\Omega$)

7.2.2 The Practical Characteristics of the LM741 Operational Amplifier

The Practical Characteristics of the LM741 Operational Amplifier:

- Very High Voltage Gain ($\Delta V = 200,000$)
- Very High Bandwidth ($Frequency \leq 1.5Mhz$)
- Very High Input Impedance ($Z_{in} \approx 2M\Omega$)
- Very Low Output Impedance ($Z_{out} \leq 75\Omega$)

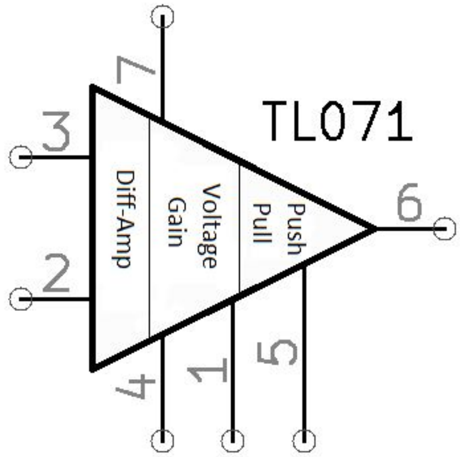


Figure 7.2: Operational Amplifier Basic Internal Arrangement

7.2.3 Basic Internal Arrangement of the Operational Amplifier

The basic internal arrangement of an Operational Amplifier consists of an input Differential Amplifier stage which provides the two out-of-phase inputs, the plus and minus inputs. The next stage is the Voltage Gain stage which provides all the necessary gain. The final stage of the Operational Amplifier is the Push-Pull Amplifier, which provides a current gain capable of up to approximately 50mA. Additionally, the operational amplifier has built-in output short-circuit protection allowing the output to be shorted.

7.2.4 Modes of Operation

Negative Feedback is a highly valuable concept in operational amplifier applications. It involves returning a portion of the amplifier's output voltage to the input with a phase angle that opposes or subtracts from the input signal.

Differential Mode

In Differential Mode the Operational Amplifier will have a signal applied to one input with the other input grounded OR two 180° (opposite) polarity signals are applied to the inputs.

Single-Ended Differential Mode:

- **Non-Inverting, Single-Ended Differential Mode** Signal is applied to the Non-Inverting Input. The Inverting Input is connected to the reference.
- **Inverting, Single-Ended Differential Mode** Signal is applied to the Inverting Input. The Non-Inverting Input is connected to the reference.

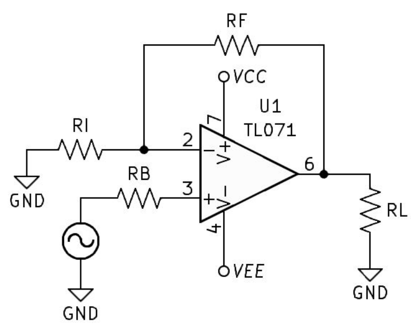


Figure 7.3: Non-Inverting, Single-Ended Differential Mode

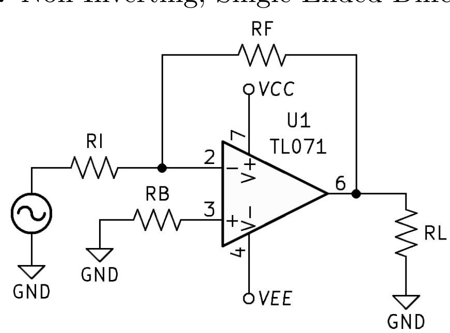


Figure 7.4: Inverting, Single-Ended Differential Mode

Double-Ended Differential Mode:

- **Differential Mode** Inputs are 180° out of phase.
- **Common Mode** Inputs are in-phase.

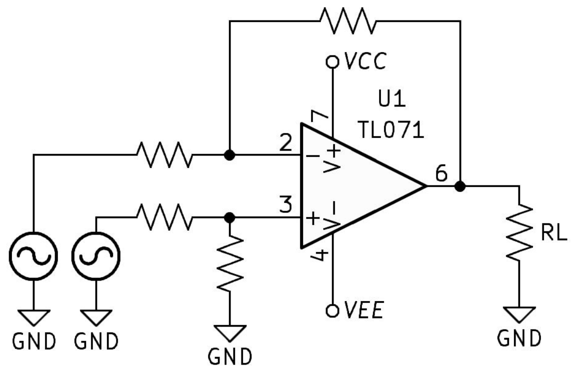


Figure 7.5: Differential Mode, Inputs are 180° out of phase

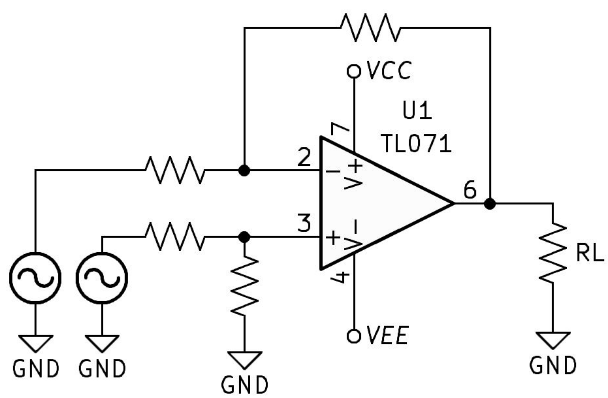


Figure 7.6: Common Mode, Inputs are in phase

7.2.5 Op-Amp Rules:

1. The output will do whatever it can to make the two input voltages equal.
2. No signal current will flow into or out of the inputs.

7.2.6 Op-Amp Steps:

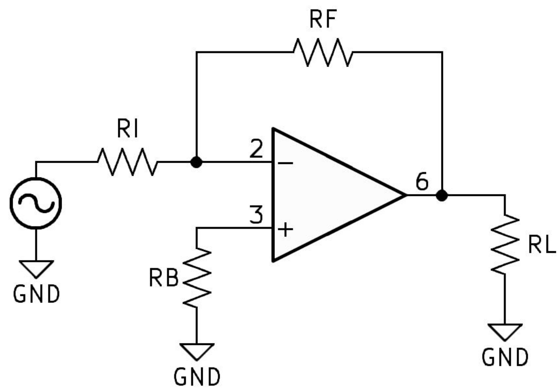


Figure 7.7: Inverting Operational Amplifier

1. Find $V_{(+)}$.
2. With feedback, $V_{(-)} = V_{(+)}$.
3. Find V_{RI}.
4. Find I_{RI}.
5. Find V_{RF}.
6. Kirchhoff to find V_{RL}.

7.2.7 Biasing Resistor (RB):

The two inputs of an Operational Amplifier should see the same or equivalent impedance. The op-amp will operate with mismatched impedances between the inputs however, this is not ideal and may cause DC offset at the output.

For example Figure 7.7.

$RB = RI//RF$

7.3 Operational Amplifier Specification Information

7.3.1 Terminology

Δ_{OL} = Open Loop Gain

Δ_{CM} = Common Mode Gain

Common Mode Rejection Ration:

$$CMRR = \frac{\Delta_{OL}}{\Delta_{CM}}$$

$$CMRR_{dB} = 20log\frac{\Delta_{OL}}{\Delta_{CM}}\ (dB)$$

Slew Rate is the maximum rate of output voltage change in response to an instantaneous change in input voltage. Slew Rate is often expressed in data sheets as volts per microsecond.

$$SlewRate_{LM741} = \frac{0.5V}{1\mu S}$$

TEXAS INSTRUMENTS

LM741

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6.5 Electrical Characteristics, LM741[(1)]

PARAMETER	TEST CONDITIONS		MIN	TYP	MAX	UNIT
Input offset voltage	$R_S \le 10$ kΩ	T_A = 25°C		1	5	mV
		$T_{AMIN} \le T_A \le T_{AMAX}$			6	mV
Input offset voltage adjustment range	T_A = 25°C, V_S = ±20 V			±15		mV
Input offset current	T_A = 25°C			20	200	nA
	$T_{AMIN} \le T_A \le T_{AMAX}$			85	500	
Input bias current	T_A = 25°C			80	500	nA
	$T_{AMIN} \le T_A \le T_{AMAX}$				1.5	μA
Input resistance	T_A = 25°C, V_S = ±20 V		0.3	2		MΩ
Input voltage range	$T_{AMIN} \le T_A \le T_{AMAX}$		±12	±13		V
Large signal voltage gain	V_S = ±15 V, V_O = ±10 V, $R_L \ge 2$ kΩ	T_A = 25°C	50	200		V/mV
		$T_{AMIN} \le T_A \le T_{AMAX}$	25			
Output voltage swing	V_S = ±15 V	$R_L \ge 10$ kΩ	±12	±14		V
		$R_L \ge 2$ kΩ	±10	±13		
Output short circuit current	T_A = 25°C			25		mA
Common-mode rejection ratio	$R_S \le 10$ Ω, V_{CM} = ±12 V, $T_{AMIN} \le T_A \le T_{AMAX}$		80	95		dB
Supply voltage rejection ratio	V_S = ±20 V to V_S = ±5 V, $R_S \le 10$ Ω, $T_{AMIN} \le T_A \le T_{AMAX}$		86	96		dB
Transient response — Rise time	T_A = 25°C, unity gain			0.3		μs
Transient response — Overshoot				5%		
Slew rate	T_A = 25°C, unity gain			0.5		V/μs
Supply current	T_A = 25°C			1.7	2.8	mA
Power consumption	V_S = ±15 V	T_A = 25°C		50	85	mW
		$T_A = T_{AMIN}$		60	100	
		$T_A = T_{AMAX}$		45	75	

Figure 7.8: LM741 Operational Amplifier Data Sheet Slew Rate

The ***Bandwidth*** of the Operational Amplifier is given at Unity Gain and is sometimes referred to as Gain Bandwidth Product. The LM741 with a voltage gain of one will have a Bandwidth of 1.5Mhz according to the specifications of Figure 7.9.

Electrical Characteristics, LM741A[(1)] (continued)

PARAMETER		TEST CONDITIONS		MIN	TYP	MAX	UNIT
Output voltage swing		$V_S = \pm 20$ V	$R_L \geq 10$ kΩ	±16			V
			$R_L \geq 2$ kΩ	±15			
Output short circuit current		$T_A = 25°C$		10	25	35	mA
		$T_{AMIN} \leq T_A \leq T_{AMAX}$		10		40	
Common-mode rejection ratio		$R_S \leq 50$ Ω, $V_{CM} = \pm 12$ V, $T_{AMIN} \leq T_A \leq T_{AMAX}$		80	95		dB
Supply voltage rejection ratio		$V_S = \pm 20$ V to $V_S = \pm 5$ V, $R_S \leq 50$ Ω, $T_{AMIN} \leq T_A \leq T_{AMAX}$		86	96		dB
Transient response	Rise time	$T_A = 25°C$, unity gain			0.25	0.8	μs
	Overshoot				6%	20%	
Bandwidth [(2)]		$T_A = 25°C$		0.437	1.5		MHz

Figure 7.9: LM741 Operational Amplifier Data Sheet Bandwidth at unity gain

7.4 Op-Amp Input and Output Impedance

7.4.1 Inverting Op-Amp Zin and Zout

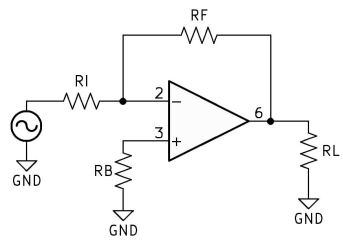


Figure 7.10: Inverting Op-Amp

$$Z_{IN(InvertingOpAmp)} = (RI + R_{Miller})//[Zin_{dev}(1 + \Delta vol(\frac{RI}{RI+RF})]$$

$R_{Miller} = \frac{RF}{1+\Delta vol}$

$LM741,\ \ Zin_{dev} = 2M\Omega$

$LM741,\ \ \Delta vol = 200,000$

$$Z_{OUT(InvertingOpAmp)} = \frac{Zout_{dev}}{1+\Delta vol(\frac{RI}{RI+RF})}$$

$LM741, Zout_{dev} = 75\Omega$

7.4.2 Non-Inverting Op-Amp Zin and Zout

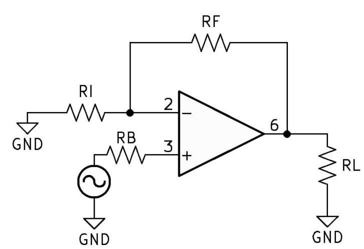


Figure 7.11: Non-Inverting Op-Amp

$$Z_{IN(NonInvertingOpAmp)} = Zin_{dev}(1 + \Delta vol(\frac{RI}{RI+RF}))$$

$LM741,\ \ Zin_{dev} = 2M\Omega$

$LM741,\ \ \Delta vol = 200,000$

$$Z_{OUT(NonInvertingOpAmp)} = \frac{Zout_{dev}}{1+\Delta vol(\frac{RI}{RI+RF})}$$

$LM741, Zout_{dev} = 75\Omega$

7.5 Op-Amp Frequency Response

Frequency Critical Low

The operational amplifier is capable of amplifying DC voltage. Low frequencies can be amplified without attenuation resulting in an Fc_{Low} equal to 0hz.

Frequency Critical High

Frequency critical high is determined largely by one of two factors in an operational amplifier. Either the Gain Bandwidth Product or the Slew Rate.

Gain Bandwidth Product sometimes referred to as FAB, is equal to the critical frequency multiplied by the voltage gain. Because the frequency critical low is essentially zero, the bandwidth of the operational amplifier is determined by the high critical frequency. The circuit gain and bandwidth have an inverse proportional relationship, as the circuit's gain increases, the frequency response or critical frequency will decrease. The data sheet will typically provide the bandwidth specification at unity gain.

$$GBP = \Delta V \times BW$$

$$GBP = \Delta V \times Fc_{High}$$

$$Fc_{High} = \frac{GBP}{\Delta V}$$

The ***Slew Rate*** and output peak voltage will also impact the bandwidth of the operational amplifier. Both are used in the following formula to determine the critical frequency high.

$$Fc_{High} = \frac{SlewRate}{2\pi vp}$$

We now have two formulas representing Fc_{High}. The actual predicted Fc_{High} will be the smaller of the two calculated formulas.

$Fc_{High} = \frac{GBP}{\Delta V}$ OR $Fc_{High} = \frac{SlewRate}{2\pi vp}$ (The smaller of the two will be Fc_{High}!)

7.6 Operational Amplifiers Active Integration and Differentiation

7.6.1 Active Integration

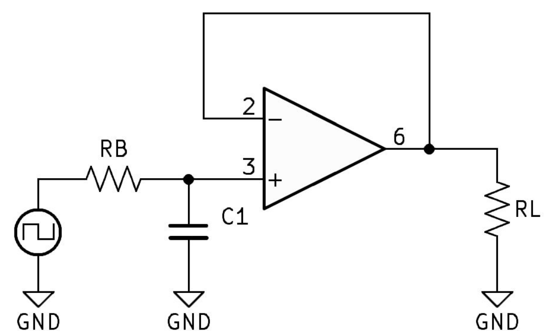


Figure 7.12: Active Integration Circuit

In figure 7.12 the Operational Amplifier is used as a buffer. Connecting the load directly to the integrator would alter the characteristics of the integrator, therefore it is necessary to connect the Op Amp as a buffer to provide circuit isolation between the load and the integrator circuit. Review RC Integration Section 2.10 on page 61.

- Op-Amp is being used as a buffer

7.6.2 Active Integration with Gain

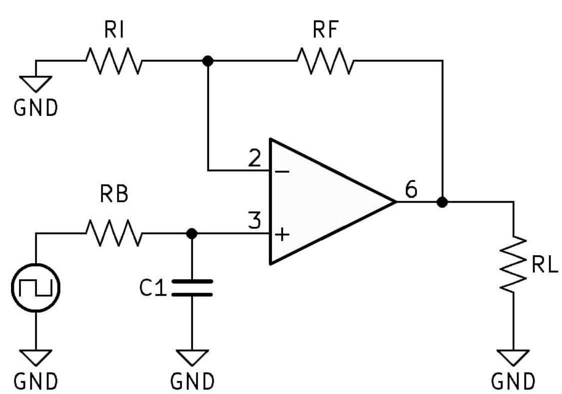


Figure 7.13: Active Integration with Gain Circuit

In addition to circuit isolation (buffer), the Op-Amp can also be used to provide signal gain.

- $\Delta V = \frac{RF}{RI}$

7.6.3 Active Differentiation

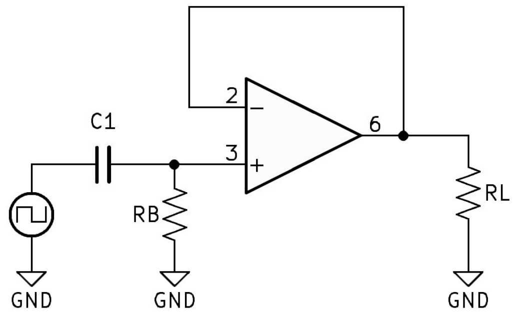


Figure 7.14: Active Differentiation Circuit

In figure 7.14 the Operational Amplifier is used as a buffer. Connecting the load directly to the differentiator would alter the characteristics of the differentiator, therefore it is necessary to connect the Op Amp as a buffer to provide circuit isolation between the load and the differentiator circuit. Review RC Differentiation Section 2.11 on page 65.

- Op-Amp is being used as a buffer

7.6.4 Active Differentiation with Gain

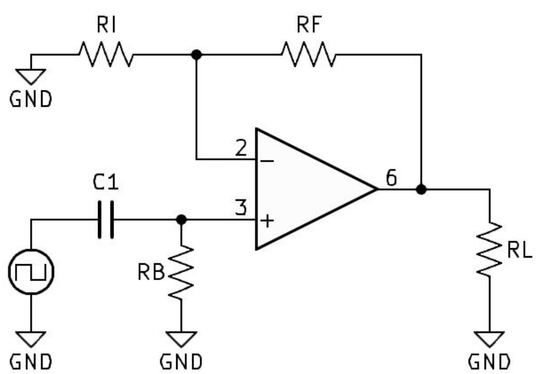


Figure 7.15: Active Differentiation with Gain Circuit

In addition to circuit isolation (buffer), the Op-Amp can also be used to provide signal gain.

- $\Delta V = \frac{RF}{RI}$

7.7 Operational Amplifier Filtering

7.7.1 First Order Active Filtering

Low-Pass Buffer

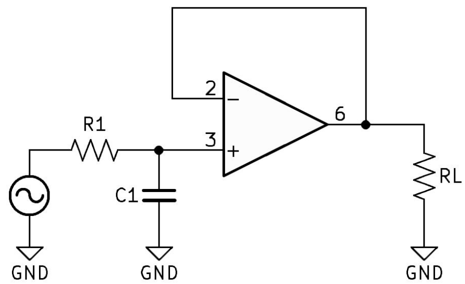


Figure 7.16: First Order Low-Pass Active Filter Buffer Circuit

In figure 7.16 the Operational Amplifier is used as a buffer. Connecting the load directly to the RC filter would alter the characteristics of the filter, therefore it is necessary to connect the Op Amp as a buffer to provide circuit isolation between the load and the filter circuit.

- Op-Amp is being used as a buffer
- Frequency roll-off is $20dB/Decade$

Low-Pass with Gain

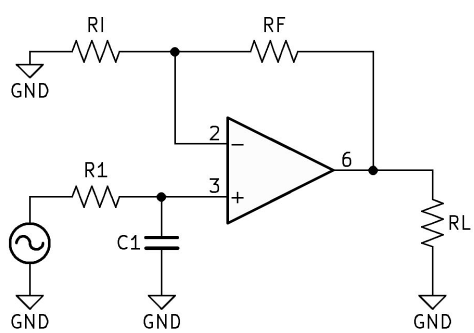


Figure 7.17: First Order Low-Pass Active Filter with Gain Circuit

In figure 7.17 the Operational Amplifier is used to provide circuit isolation and signal gain.

- $\Delta V = \frac{RF}{RI} + 1$
- Frequency roll-off is $20dB/Decade$

High-Pass Buffer

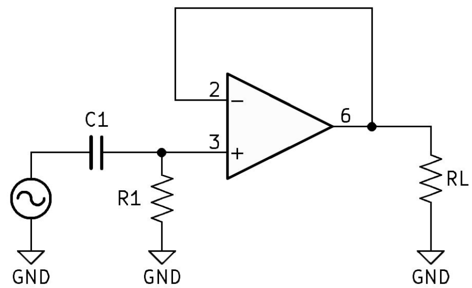


Figure 7.18: First Order High-Pass Active Filter Buffer Circuit

In figure 7.18 the Operational Amplifier is used as a buffer. Connecting the load directly to the RC filter would alter the characteristics of the filter, therefore it is necessary to connect the Op Amp as a buffer to provide circuit isolation between the load and the filter circuit.

- Op-Amp is being used as a buffer
- Frequency roll-off is $20dB/Decade$

High-Pass with Gain

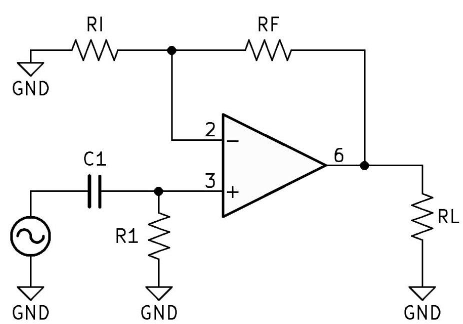


Figure 7.19: First Order High-Pass Active Filter with Gain Circuit

In figure 7.19 the Operational Amplifier is used to provide circuit isolation and signal gain.

- $\Delta V = \frac{RF}{RI} + 1$
- Frequency roll-off is $20dB/Decade$

Cascading Band-Pass Filter

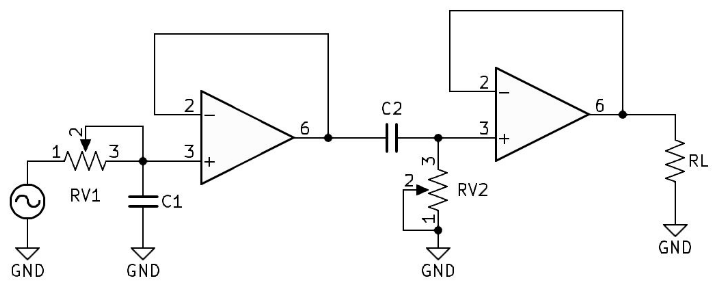


Figure 7.20: First Order Cascaded Band-Pass Active Filter

In figure 7.20 two first-order filters, a Low-Pass and a High-Pass are cascaded together to create a Band-Pass Filter.

- FC_{Low} is determined by RV1 and C1
- FC_{High} is determined by RV2 and C2
- Frequency roll-off is $20dB/Decade$

Single Stage Band-Pass Filter

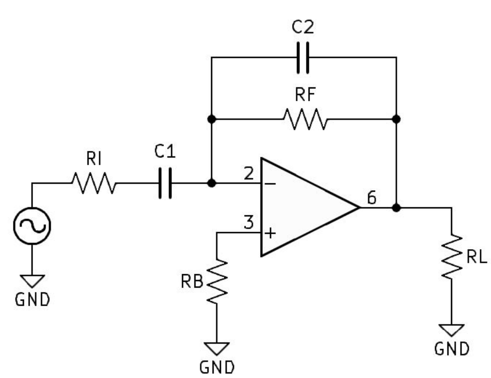


Figure 7.21: First Order Single Stage Band Pass Filter

In figure 7.21 a single-stage Operational amplifier is used to create a first-order band-pass filter.

- FC_{Low} is determined by R1 and C1, $FC_{Low} = \frac{1}{2\pi R1C1}$
- FC_{High} is determined by R2 and C2, $FC_{High} = \frac{1}{2\pi R2C2}$
- Frequency roll-off is $20dB/Decade$

7.7.2 Second Order Active Filtering

Second Order Low-Pass

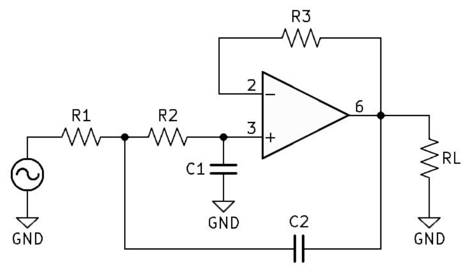


Figure 7.22: Second-Order Low Pass Active Filter

- Op-Amp is used as a buffer and active filtering
- $R1 = R2$
- $C2 = 2 \times C1$
- $FC_{High} = \frac{1}{2\pi\sqrt{R1 \times R2 \times C1 \times C2}}$
- Frequency roll-off is $40dB/Decade$

Second Order High-Pass

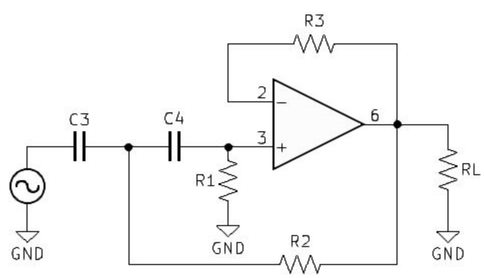


Figure 7.23: Second-Order High Pass Active Filter

- Op-Amp is used as a buffer and active filtering
- $C1 = C2$
- $R1 = 2 \times R2$
- $FC_{High} = \frac{1}{2\pi\sqrt{R1 \times R2 \times C1 \times C2}}$
- Frequency roll-off is $40dB/Decade$

Second-Order Cascading Band-Pass Filter

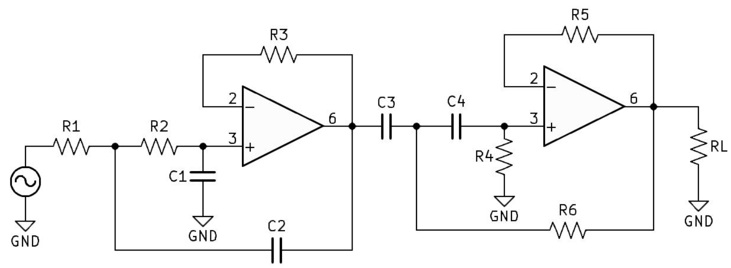


Figure 7.24: Second-Order Cascaded Band Pass Active Filter

- Two Second-Order Cascaded Op-Amp filters (Low-Pass to a High-Pass)
- Frequency roll-off is $40dB/Decade$

Second-Order Single Stage Band-Pass Filter

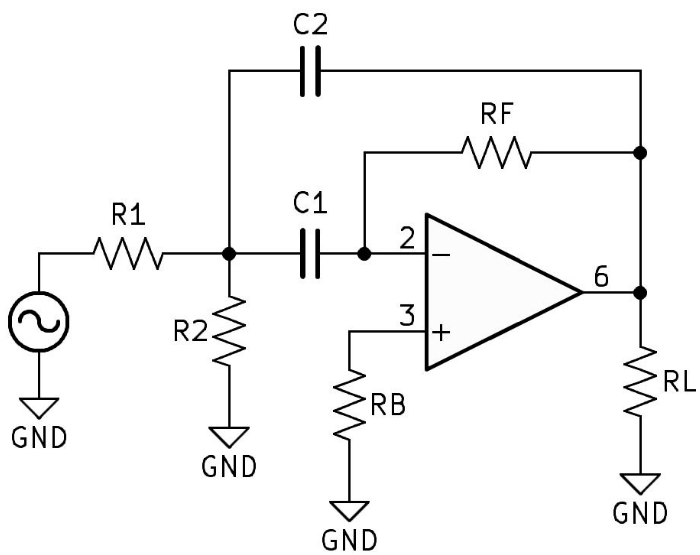


Figure 7.25: Second-Order Single Stage Band Pass Active Filter

- See Design Steps on next page

Second-Order Single Stage Band-Pass Filter Design Steps

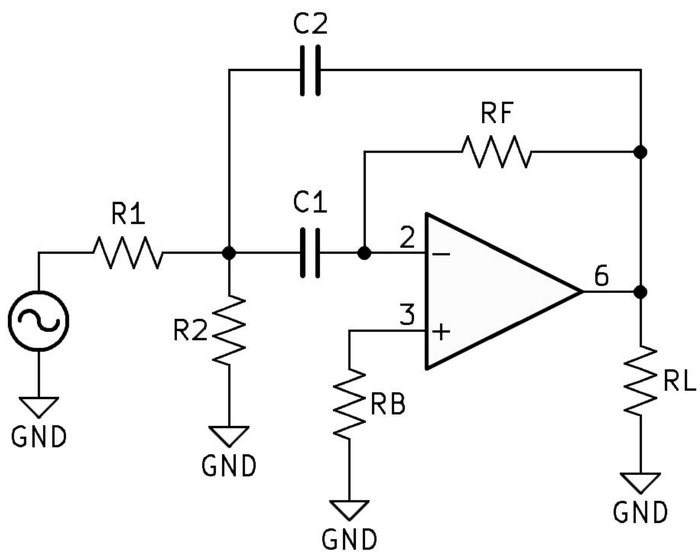


Figure 7.26: Second-Order Single Stage Band Pass Active Filter

Second-Order Single Stage Band Pass Active Filter Design Steps:

- $Q = \frac{f_R}{BW}$
- Use small caps $\approx 0.001uF$

  $C1 = C2$

- Pick:

  Resonant Frequency

  Bandwidth

  Q *(Q should be between 1 & 20 to prevent oscillations)

- Pick Δv at f_R *(make $\Delta v < 2Q^2$)
- $R1 = \frac{Q}{2\pi f_R C \Delta v}$
- $R2 = \frac{Q}{2\pi f_R C (2Q^2 - \Delta v)}$
- $RF = \frac{2Q}{2\pi f_R C}$
- $RB = RF$

7.8 Bessel, Butterworth, & Chebyshev Filters

7.8.1 Bessel Filter

- **Named after**: German Mathematician Friedrick Bessel
- **Notable features**:The Bessel filter has the gentlest response of the group. Even though it doesn't have a sharp cutoff, it offers superior phase shift (delay) compared to the other filters in the group. The Bessel filter requires the most stages (i.e. most components); however, it offers excellent characteristics: low sensitivity to component tolerance, superior step response.
- **Best used for**: The Bessel filter is ideal for applications that require minimal phase shift. Due to the gentle frequency response of the Bessel filter, it can only be used in applications where there is adequate space between the passband and stopband. - Filter Topology Face Off: A closer look at the top 4 filter types [13]

7.8.2 Butterworth Filter

- **Named after**: British Physicist Stephen Butterworth
- **Notable features**:The Butterworth filter is commonly referred to as the "maximally flat" option because the passband response offers the steepest roll-off without inducing a passband ripple. In addition to the flat passband response, the selectivity of the Butterworth filter is better than many other filter typologies such as the Bessel or Gaussian. The flip-side of this improved selectivity is greater delay and poorer phase linearity. Butterworth filters offer solid performance considering the number of components needed to implement the filter.
- **Best used for**: Butterworth filters are typically forgiving to part tolerances and values of discrete elements (capacitors, inductors, and resistors). For most bandpass designs, the (Voltage Standing Wave Ratio) VSWR at center frequency is extremely good. - Filter Topology Face Off: A closer look at the top 4 filter types [13]

7.8.3 Chebyshev Filter

- **Named after**: Russian mathematician Pafnuty Chebyshev
- **Notable features**: The Chebyshev filter is known for its ripple response. This ripple response can be designed to be present in the passband (Chebyshev Type 1) or in the stopband (Chebyshev Type 2). The amplitude of the ripple is directly proportional to the steepness of the rolloff. That is, if you want a steeper response, you'll see a larger ripple response. These filters offer performance between that of Elliptic function filters and Butterworth

filters. The phase response of the Chebyshev filter is relatively non-linear, which ultimately wreaks havoc on demodulators because it tends to distort pulses because of the non-linear delays. The most common workaround for this phenomenon is to increase the bandwidth of the Chebyshev filter to push this non-linear region further out.

- **Best used for**: The Chebyshev filter is the workhorse of the common filter typologies. Its response is easily realized with few components and offers excellent selectivity with one of the steepest roll-off responses of the group. - Filter Topology Face-Off: A closer look at the top 4 filter types [13]

7.8.4 Damping Factor

In filter design, the damping factor (ζ) is important for shaping the frequency response, controlling the bandwidth of the filter, and managing the transient response characteristics. For example, in a second-order low-pass filter, the damping factor directly affects the peak response at the cutoff frequency.

- **Overdamped** ($d > 1$) : The system returns to equilibrium without oscillating. It has a slower response but no overshoot.
- **Critically Damped** ($d = 0.707$): The system returns to equilibrium as quickly as possible without oscillating.
- **Underdamped** ($d < 1$): The system will start to oscillate especially as the gain is increased.

Most designs of second-order filters are generally named after their inventor with the most common filter types being: Butterworth, Chebyshev, Bessel and Sallen-Key. All these types of filter designs are available as either: low pass filter, high pass filter, band pass filter and band stop (notch) filter configurations, and being second order filters, all have a 40-dB-per-decade roll-off.

In active second-order filters, the damping factor, $\zeta(zeta)$, which is the inverse of Q is normally used. Both Q and ζ are independently determined by the gain of the amplifier, A so as Q decreases the damping factor increases. In simple terms, a low pass filter will always be low pass in its nature but can exhibit a resonant peak in the vicinity of the cut-off frequency, that is the gain can increase rapidly due to resonance effects of the amplifier's gain.

Then Q, the quality factor, represents the "peakiness" of this resonance peak, that is its height and narrowness around the cut-off frequency point, f_C. But a filters gain also determines the amount of its feedback and therefore has a significant effect on the frequency response of the filter.

Generally to maintain stability, an active filter's gain must not be more than 3.

Then somewhere in between, $\zeta = 0$ and $\zeta = 2.0$, there must be a point where the frequency response is of the correct value, and there is. This is when the filter is "critically damped" and occurs when $\zeta = 0.7071$. - Second Order Filters[14]

7.8.5 Second Order Low-Pass Test Circuit

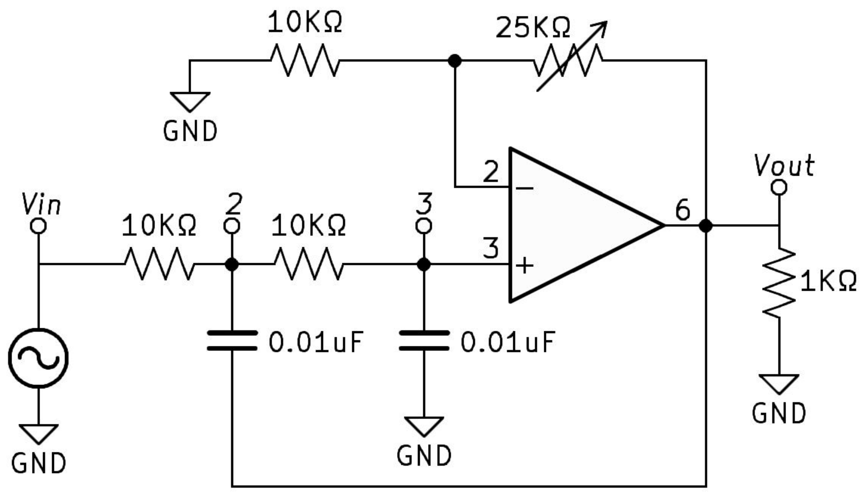


Figure 7.27: Bessel, Butterworth, & Chebyshev Test Circuit

Table 7.1: Bessel, Butterworth, & Chebyshev Test Circuit Data

	Vin	2	3	ζ	RV	ΔV	Vout
Bessel	2vpp	1.414vpp	1vpp	2	0.0Ω	1	1vpp
Bessel	1.73vpp	1.414vpp	1vpp	1.73	2.68KΩ	1.268	1.268vpp
Butterworth	1.414vpp	1.414vpp	1vpp	1.414	5.86KΩ	1.586	1.586vpp
Chebyshev	1vpp	1.414vpp	1vpp	1	10KΩ	2	2vpp
Chebyshev	0.5vpp	1.414vpp	1vpp	0.5	15KΩ	2.5	2.5vpp
Oscillator	0.0vpp	Undetermined		0	≥20KΩ	≥ 3	OSC.

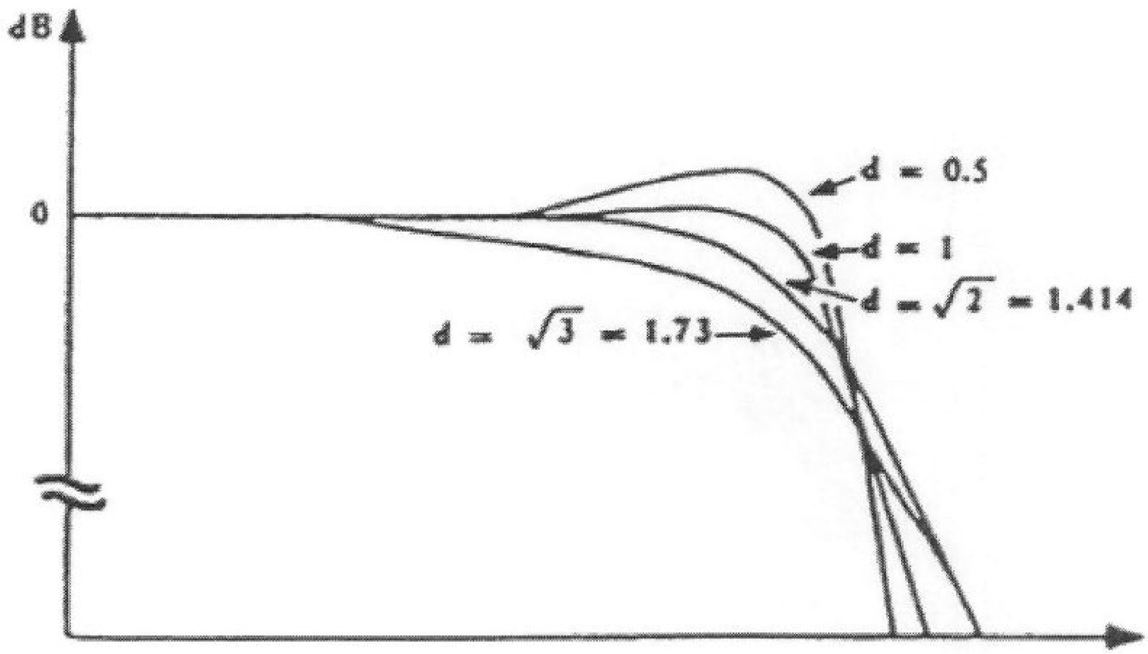
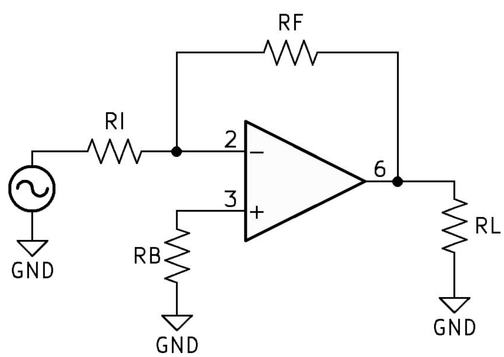


Figure 7.28: Filter Response Curves in terms of Damping, d = zeta(ζ)

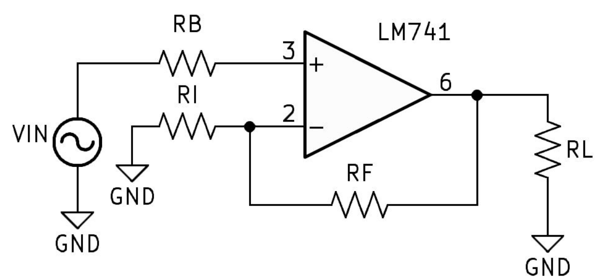
7.9 Operational Amplifier Practice Problems

1. **Given:** $RF = 10K\Omega, RI = 1K\Omega, Slew\ Rate = 0.5V/\mu S, FAB = 1.5MHz, \&\ VIN = 1vp$.
   Find: $Vout$, ΔV, RB, Z_{IN}, Z_{OUT}, and FC_{High}.



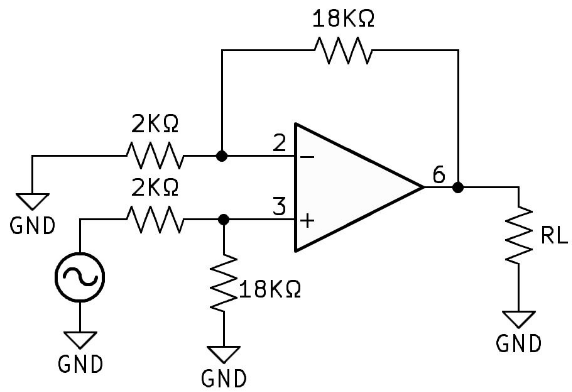
$Vout =$	$\Delta V =$	$RB =$
$Z_{IN} =$	$Z_{OUT} =$	$FC_{High} =$

2. **Given:** $RF = 10K\Omega, RI = 1K\Omega, Slew\ Rate = 0.5V/\mu S, FAB = 1.5MHz, \&\ VIN = 1vp$.
   Find: $Vout$, ΔV, RB, Z_{IN}, Z_{OUT}, and FC_{High}.



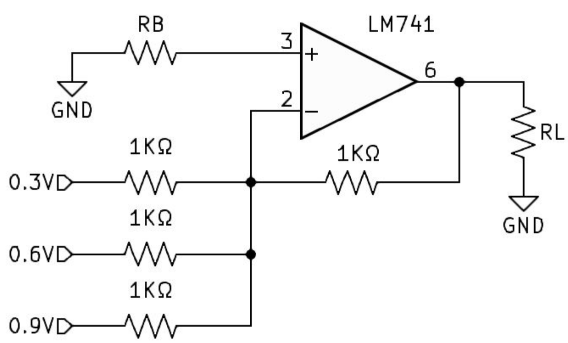
$Vout =$	$\Delta V =$	$RB =$
$Z_{IN} =$	$Z_{OUT} =$	$FC_{High} =$

3. **Given:** $VIN = 250mvp$.
   Find: $Vout$ and ΔV.



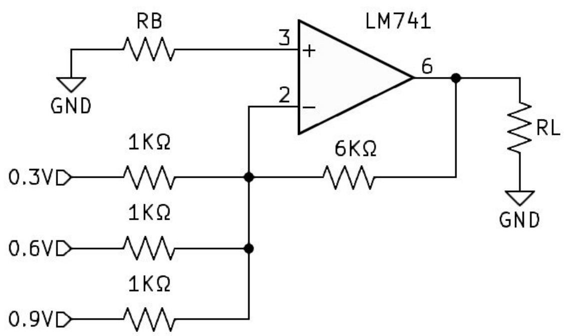
$Vout =$	$\Delta V =$

4. **Given:** See Circuit Schematic
   Find: $Vout$ and RB.



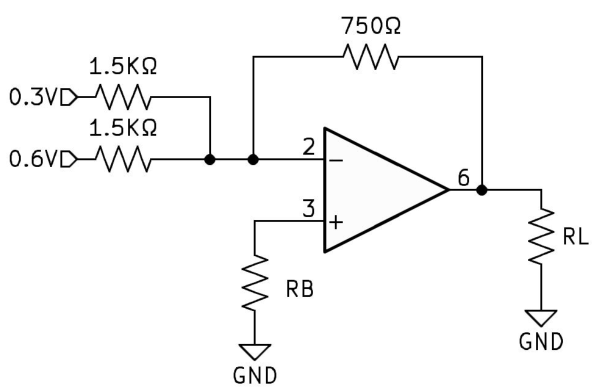
$Vout =$	$RB =$

5. **Given:** See Circuit Schematic
   Find: $Vout$ and RB.



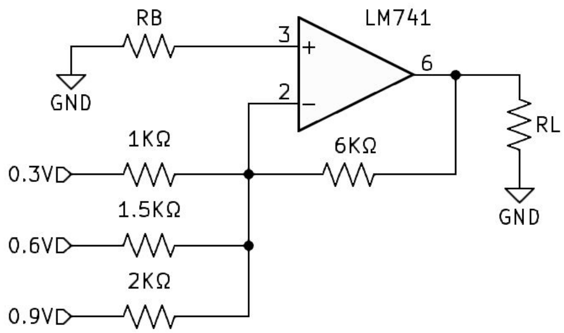
$Vout =$	$RB =$

6. **Given:** See Circuit Schematic
   Find: $Vout$ and RB.



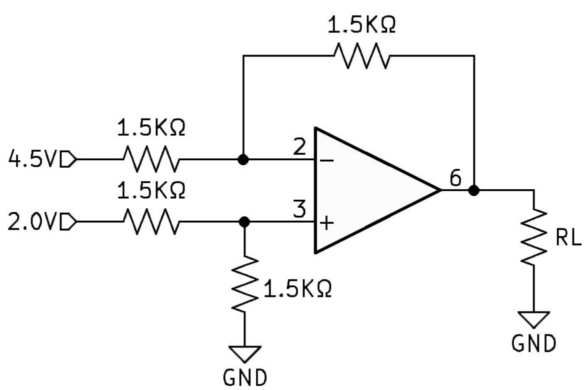
$Vout =$	$RB =$

7. **Given:** See Circuit Schematic
   Find: $Vout$ and RB.



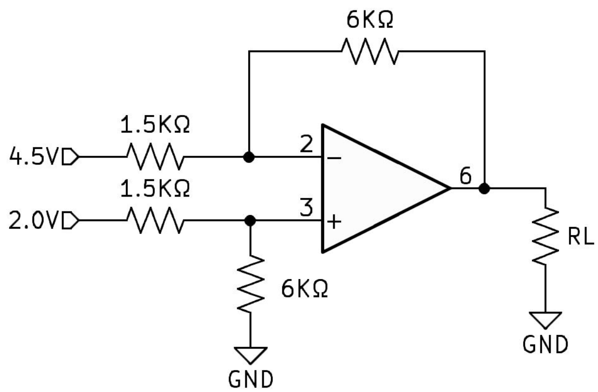
$Vout =$	$RB =$

8. **Given:** See Circuit Schematic
   Find: $Vout$.



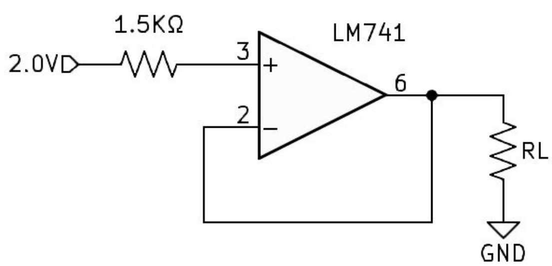
$Vout =$

9. **Given:** See Circuit Schematic
   Find: $Vout$.



$Vout =$

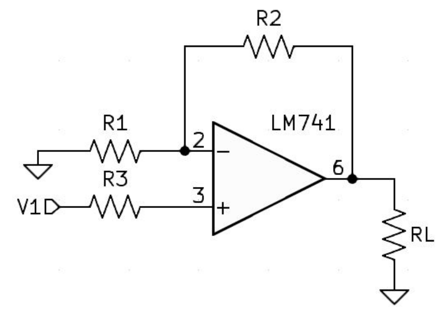
10. **Given:** See Circuit Schematic
    Find: $Vout$.



$Vout =$

11. **Given:** $V_{Supply} = \pm 15V$, $Zin_{Dev} = 2M\Omega$, $Zout_{Dev} = 75\Omega$, $Slew_{Rate} = 1.5V/\mu S$, $GBWP = 1.5MHz$, $R1 = 1K\Omega$, $R2 = 10K\Omega$, & $V1 = 600mV$

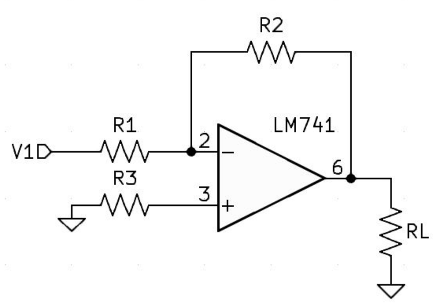
    Find: $R3$, V_{RL}, ΔV, Z_{IN}, Z_{OUT}, & FC_{High}



$R3 =$	$V_{RL} =$	$\Delta V =$
$Z_{IN} =$	$Z_{OUT} =$	$FC_{High} =$

12. **Given:** $V_{Supply} = \pm 25V$, $Zin_{Dev} = 2.5M\Omega$, $Zout_{Dev} = 65\Omega$, $Slew_{Rate} = 0.5V/\mu S$, $GBWP = 1.0MHz$, $R1 = 4.5K\Omega$, $R2 = 7.4K\Omega$, & $V1 = 10V$

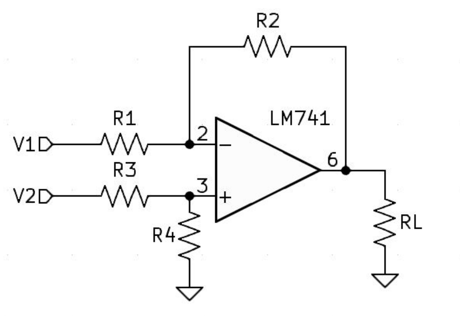
    Find: $R3$, V_{RL}, ΔV, Z_{IN}, Z_{OUT}, & FC_{High}



$R3 =$	$V_{RL} =$	$\Delta V =$
$Z_{IN} =$	$Z_{OUT} =$	$FC_{High} =$

13. **Given:** $V_{Supply} = \pm 15V$, $R1 = 420\Omega$, $R2 = 4.2K\Omega$, $R3 = 420\Omega$, $R4 = 4.2K\Omega$, $V1 = 200mV$, & $V2 = 500mV$

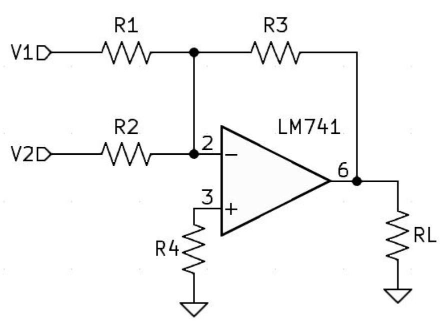
    Find: V_{RL}



$V_{RL} =$

14. **Given:** $V_{Supply} = \pm 15V$, $R1 = 1K\Omega$, $R2 = 1K\Omega$, $R3 = 1K\Omega$, $V1 = 4V$, & $V2 = 4V$

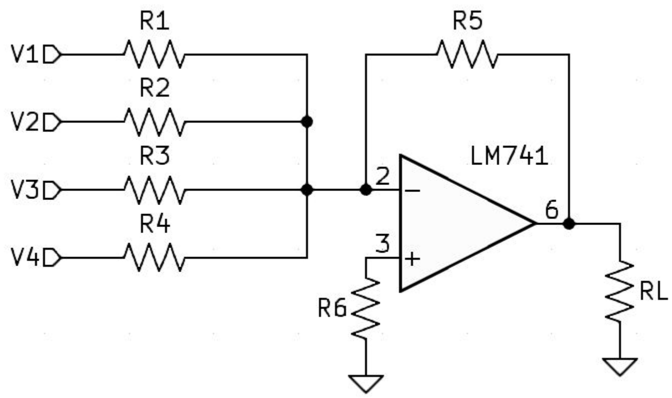
    Find: V_{RL} & $R4$



$V_{RL} =$	$R4 =$

15. **Given:** $V_{Supply} = \pm18V$, $R1 = 50K\Omega$, $R2 = 25K\Omega$, $R3 = 12.5K\Omega$, $R4 = 6.25K\Omega$, $R5 = 10K\Omega$, & $V1, V2, V3, \& \ V4 = 5V$

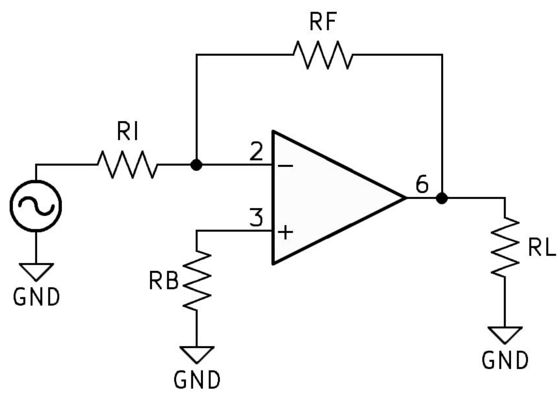
    Find: V_{RL} & $R6$



$V_{RL} =$	$R6 =$

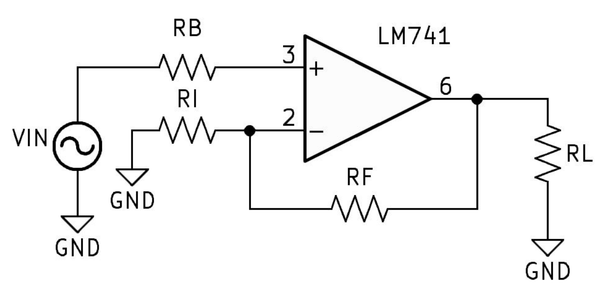
7.10 Operational Amplifier Practice Problem Answers

1. **Given:** $RF = 10K\Omega, RI = 1K\Omega, Slew\ Rate = 0.5V/\mu S, FAB = 1.5MHz, \&\ VIN = 1vp.$
   Find: $Vout$, ΔV, RB, Z_{IN}, Z_{OUT}, and FC_{High}.



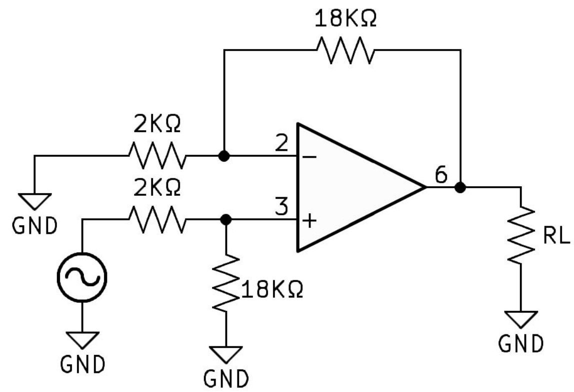
$Vout = 10v_P$	$\Delta V = -10$	$RB = 909.1\Omega$
$Z_{IN} = 1K\Omega$	$Z_{OUT} = 4.125m\Omega$	$FC_{High} = 7.958Khz$

2. **Given:** $RF = 10K\Omega, RI = 1K\Omega, Slew\ Rate = 0.5V/\mu S, FAB = 1.5MHz, \&\ VIN = 1vp.$
   Find: $Vout$, ΔV, RB, Z_{IN}, Z_{OUT}, and FC_{High}.



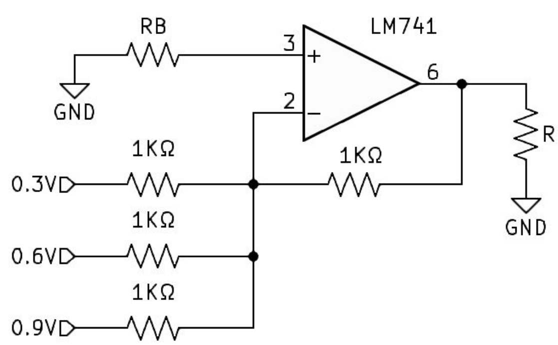
$Vout = 11v_P$	$\Delta V = 11$	$RB = 909.1\Omega$
$Z_{IN} = 36.364 \times 10^9\ \Omega$	$Z_{OUT} = 4.125m\Omega$	$FC_{High} = 7.234Khz$

3. **Given:** $VIN = 250mvp$.
   Find: $Vout$ and ΔV.



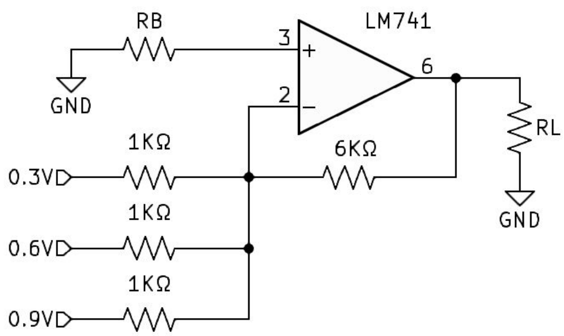
$Vout = 2.25v_P$	$\Delta V = 9$

4. **Given:** See Circuit Schematic
   Find: $Vout$ and RB.



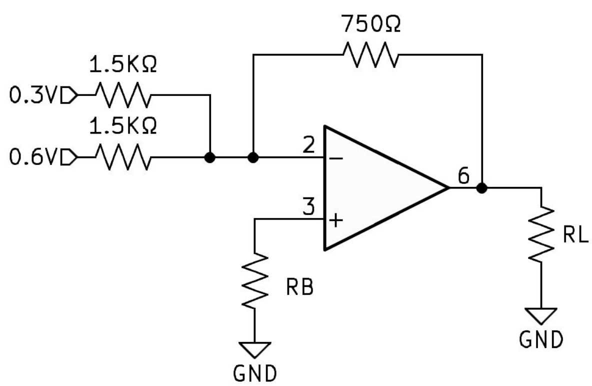
$Vout = -1.8V$	$RB = 250\Omega$

5. **Given:** See Circuit Schematic
   Find: $Vout$ and RB.



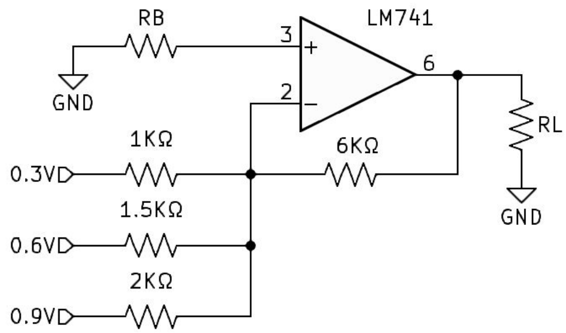
$Vout = -10.8V$	$RB = 316.8\Omega$

6. **Given:** See Circuit Schematic
   Find: $Vout$ and RB.



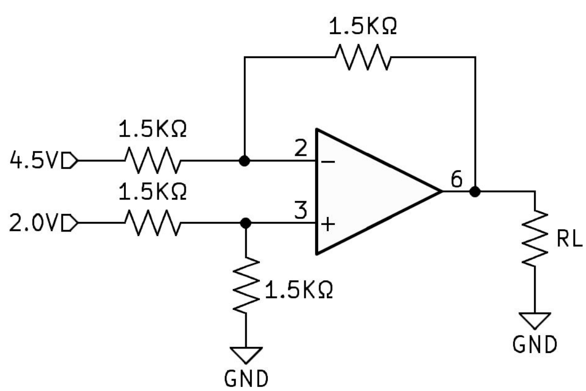
$Vout = -450mV$	$RB = 375\Omega$

7. **Given:** See Circuit Schematic
   Find: $Vout$ and RB.



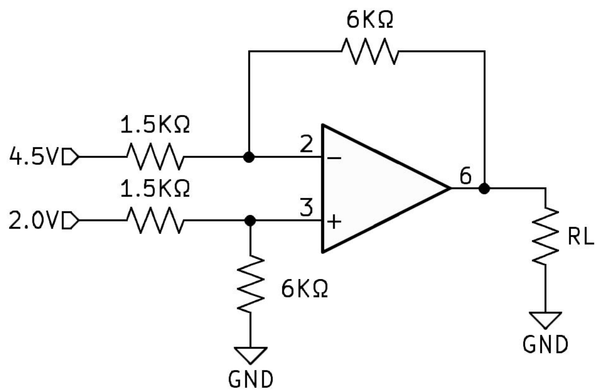
$Vout = -6.9V$	$RB = 429\Omega$

8. **Given:** See Circuit Schematic
   Find: $Vout$.



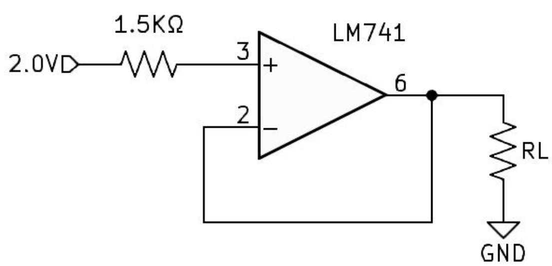
$Vout = -2.5V$

9. **Given:** See Circuit Schematic
   Find: $Vout$.



$Vout = -10V$

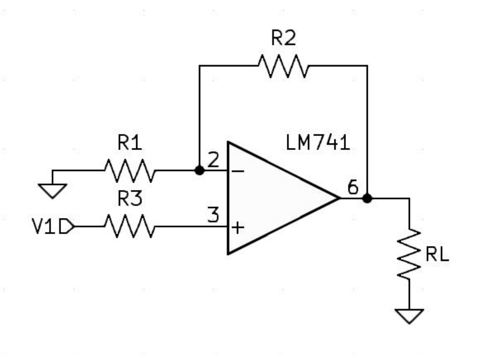
10. **Given:** See Circuit Schematic
    Find: $Vout$.



$Vout = 2V$

11. **Given:** $V_{Supply} = \pm 15V$, $Zin_{Dev} = 2M\Omega$, $Zout_{Dev} = 75\Omega$, $Slew_{Rate} = 1.5V/\mu S$, $GBWP = 1.5MHz$, $R1 = 1K\Omega$, $R2 = 10K\Omega$, & $V1 = 600mV$

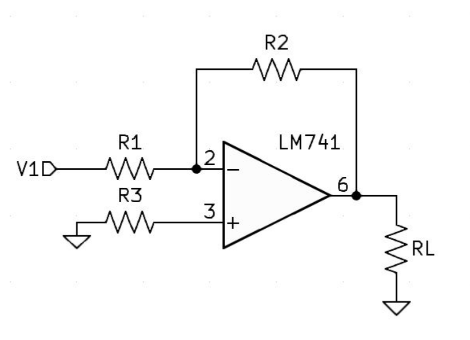
    Find: $R3$, V_{RL}, ΔV, Z_{IN}, Z_{OUT}, & FC_{High}



$R3 = 909.9\Omega$	$V_{RL} = 6.6V$	$\Delta V = 11$
$Z_{IN} = 36.189G\Omega$	$Z_{OUT} = 4.125m\Omega$	$FC_{High} = 36.173KHz$

12. **Given:** $V_{Supply} = \pm 25V$, $Zin_{Dev} = 2.5M\Omega$, $Zout_{Dev} = 65\Omega$, $Slew_{Rate} = 0.5V/\mu S$, $GBWP = 1.0MHz$, $R1 = 4.5K\Omega$, $R2 = 7.4K\Omega$, & $V1 = 10V$

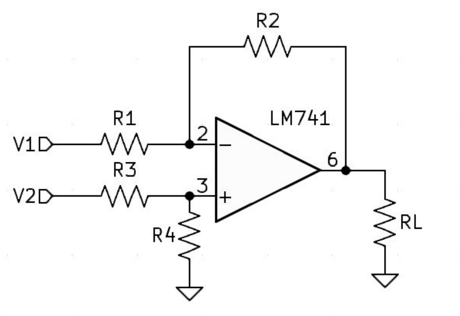
    Find: $R3$, V_{RL}, ΔV, Z_{IN}, Z_{OUT}, & FC_{High}



$R3 = 2.798K\Omega$	$V_{RL} = -16.444V$	$\Delta V = -1.644$
$Z_{IN} = 4.5K\Omega$	$Z_{OUT} = 859.433\mu\Omega$	$FC_{High} = 4.839KHz$

13. **Given:** $V_{Supply} = \pm 15V$, $R1 = 420\Omega$, $R2 = 4.2K\Omega$, $R3 = 420\Omega$, $R4 = 4.2K\Omega$, $V1 = 200mV$, & $V2 = 500mV$

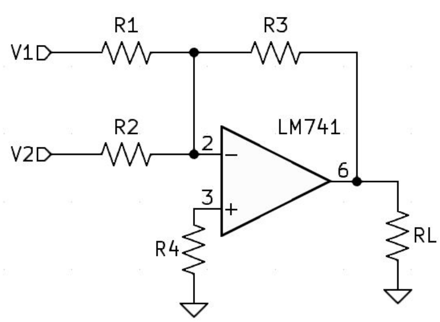
    Find: V_{RL}



$V_{RL} = 3V$

14. **Given:** $V_{Supply} = \pm 15V$, $R1 = 1K\Omega$, $R2 = 1K\Omega$, $R3 = 1K\Omega$, $V1 = 4V$, & $V2 = 4V$

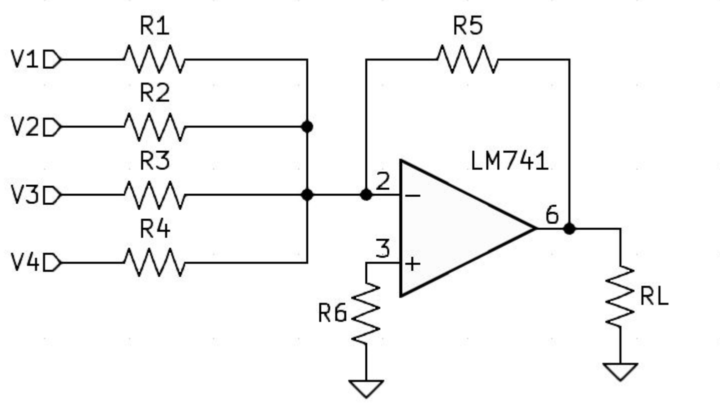
    Find: V_{RL} & $R4$



$V_{RL} = -8V$	$R4 = 333.333\Omega$

15. **Given:** $V_{Supply} = \pm 18V$, $R1 = 50K\Omega$, $R2 = 25K\Omega$, $R3 = 12.5K\Omega$, $R4 = 6.25K\Omega$, $R5 = 10K\Omega$, & $V1, V2, V3, \& \ V4 = 5V$

Find: V_{RL} & $R6$



$V_{RL} = -15V$	$R6 = 2.5K\Omega$

Week 8

Digital to Analog & Analog to Digital Conversions

8.1 Objectives

8.1.1 Digital to Analog Conversion Objectives

Knowledge Objectives

1. **Understand DAC Fundamentals:** Explain the basic principles of digital-to-analog conversion, including key concepts such as resolution, sampling rate, and quantization.
2. **Types of DACs:** Describe the different types of DACs (e.g., R-2R ladder & binary-weighted) and their operating principles.
3. **DAC Specifications:** DAC Specifications: Understand and interpret specifications of DACs, including linearity, and settling time.

Design Objectives

1. **Design a Simple DAC Circuit:** Design and implement a basic DAC circuit using common components.
2. **Component Selection:** Select appropriate components for DAC circuits based on application requirements, such as resolution and speed.
3. **Circuit Pre-Calculations:** Use calculations to model DAC behavior and predict performance characteristics and output waveforms.

Troubleshooting Objectives

1. **Identify Common Issues:** Identify and diagnose common issues in DAC circuits, such as non-linearity, glitches, and noise.

2. **Testing and Measurement:** Use oscilloscopes, signal generators, and other test equipment to measure and analyze DAC output signals.

3. **Performance Optimization:** Implement techniques to optimize DAC performance, including filtering and error correction methods.

8.1.2 Analog to Digital Conversion Objectives

Knowledge Objectives

1. **Understand ADC Fundamentals:** Explain the basic principles of analog-to-digital conversion, including key concepts such as sampling theorem, Nyquist rate, and aliasing.

2. **Types of ADCs:** Describe the different types of ADCs (e.g., flash, successive approximation, delta-sigma) and their operating principles.

3. **ADC Specifications:** Understand and interpret specifications of ADCs, including resolution, signal-to-noise ratio (SNR), and effective number of bits (ENOB).

Design Objectives

1. **Design a Simple ADC Circuit:** Design and implement a basic ADC circuit using common components.

2. **Component Selection:** Select appropriate components for ADC circuits based on application requirements, such as sample rate.

3. **Circuit Pre-Calculations:** Use calculations to model ADC behavior and predict performance characteristics.

Troubleshooting Objectives

1. **Identify Common Issues:** Identify and diagnose common issues in ADC circuits, such as quantization noise, jitter, and distortion.

2. **Testing and Measurement:** Use oscilloscopes, signal generators, and other test equipment to measure and analyze ADC input signals and conversion accuracy.

3. **Performance Optimization:** Implement techniques to optimize ADC performance, including proper grounding, shielding, and anti-aliasing filtering.

8.1.3 Integrated Objectives

Practical Applications

1. **Real-world Applications:** Discuss and analyze real-world applications of DACs and ADCs in systems such as audio processing, instrumentation, and communication systems.

2. **Integration in Systems:** Understand the integration of DACs and ADCs within larger systems, including interfacing with microcontrollers and digital signal processors (DSPs).

Project-Based Learning

1. **Lab Project:** Design, implement, and troubleshoot a complete system that incorporates both DAC and ADC circuits, demonstrating an understanding of both digital-to-analog and analog-to-digital conversion processes.

8.2 Key Terms

Acquisition Acquisition refers to the process of capturing and converting an analog signal into a digital form so it can be processed, analyzed, or stored by digital systems.

ADC Analog-to-Digital Converter. This conversion is crucial in digital electronics because most modern electronic devices process and store data in digital form.

ADC Resolution The resolution of an ADC is the number of distinct digital output values it can produce for a given analog input range. It is usually specified in bits.

An n-bit ADC can represent 2^n discrete levels. For example, a 10-bit ADC can represent $2^{10} = 1024$ distinct levels.

If an ADC has a 10-bit resolution and an input voltage range of 0 to 5 volts, each digital step represents $\frac{5V}{1024} \approx 4.88mV/step$.

Aliasing Aliasing occurs when a signal is sampled at a rate that is too low to capture its variations accurately, leading to distortion and loss of information.

Anti-Aliasing Filter A low-pass filter, applied to the analog signal before sampling, that removes frequency components higher than half the sampling rate. This ensures that the sampled signal accurately represents the original signal without high-frequency components that could cause aliasing.

DAC Digital-to-Analog Converter. A DAC is a device that converts a digital signal, which is a discrete signal represented by binary numbers, into an analog signal, which is a continuous signal. This conversion is essential in various applications where digital data needs to be converted back into a form that can be perceived by human senses or processed by analog systems.

DAC Resolution The resolution of a DAC is the number of discrete output levels it can produce from a given range of digital input values. It is also specified in bits.

An n-bit DAC can produce 2^n discrete output levels. For example, a 8-bit DAC can represent $2^8 = 256$ distinct levels.

If a DAC has an 8-bit resolution and an output voltage range of 0 to 10 volts, each step corresponds to $\frac{10V}{256} \approx 39.06mV/step$.

Delta-Sigma ADC The Delta-Sigma ADC offers high-resolution performance while providing high stability. -Texas Instruments [15]

Encoding The process of converting the quantized values into a binary code that can be processed by digital systems.

Filtering Smoothing out the step-like output of the DAC to produce a more natural analog signal, often using a low-pass filter.

Latency Latency refers to the time delay between the application of an analog input signal and the availability of the corresponding digital output. This delay encompasses all the internal processes of the ADC, including sampling, conversion, and any additional processing stages.

Nyquist theorem also known as the Nyquist-Shannon sampling theorem, is a fundamental principle in the field of digital signal processing. It provides a criterion for the minimum sampling rate required to accurately capture and reconstruct a continuous signal from its samples without introducing aliasing.

Mathematical Expression: If f_{max} is the highest frequency component of the signal, the sampling rate f_s must satisfy:

$f_s \geq 2 \times f_{max}$

Pipeline ADC Also known as subranging quantizers, pipeline ADCs consist of numerous consecutive stages, each containing a track/hold (T/H), a low-resolution ADC and DAC, and a summing circuit that includes an interstage amplifier to provide gain. -Maxim Integrated [15]

Quantization The process of mapping the sampled values to a finite set of levels, which are represented by binary numbers.

Reconstruction The process of converting discrete digital values into a continuous analog waveform.

Sampling The process of measuring the amplitude of an analog signal at regular intervals is known as the sampling rate.

SAR ADC The success of approximation register converter, or SAR ADC, is often considered the backbone of general-purpose mixed-signal circuits and is used in many data acquisition applications, including control loops, power monitoring, and low-to-medium frequency analysis. -Texas Instruments [15]

Settling time in electronics refers to the time required for a circuit or system to reach a stable state after a perturbation, such as a change in input signal or conditions. It is particularly important in analog and digital systems where signals need to stabilize to ensure accurate and reliable operation.

8.3 Digital to Analog Conversion

8.3.1 Binary Weighted DAC

Binary weighted digital-to-analog converters are a type of data converter that converts a digital binary number into an equivalent analog output signal proportional to the value of the digital number - Electronics Tutorials [16]

8.3.2 BCD Binary Weighted DAC Design

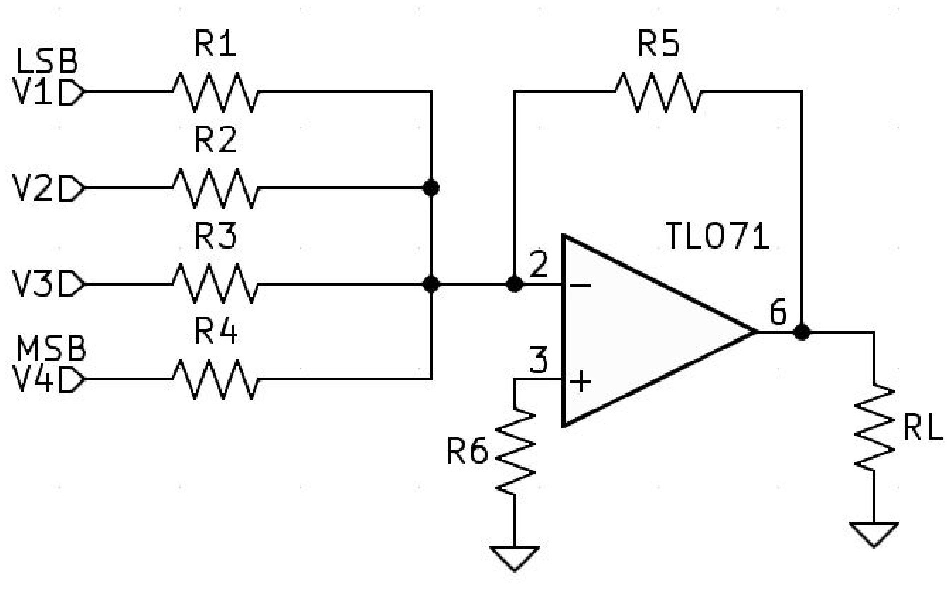


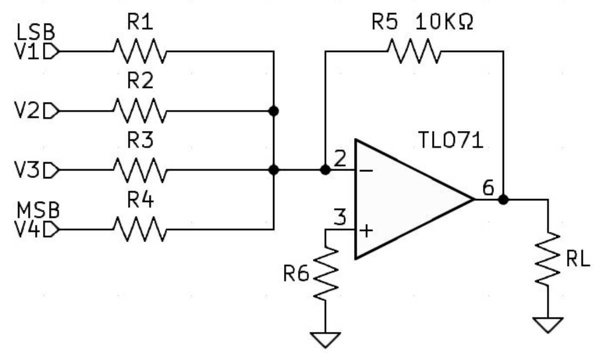
Figure 8.1: Binary Weighted Digital to Analog Converter

Table 8.1: Binary Weighted DAC Data

Digital				Analog
$(MSB)\ 2^3$	2^2	2^1	$(LSB)\ 2^0$	V_{RL}
0	0	0	0	$0V$
0	0	0	1	$-1V$
0	0	1	0	$-2V$
0	0	1	1	$-3V$
0	1	0	0	$-4V$
0	1	0	1	$-5V$
0	1	1	0	$-6V$
0	1	1	1	$-7V$
1	0	0	0	$-8V$
1	0	0	1	$-9V$

Steps:

1. See figure 8.1: Binary Weighted Digital to Analog Converter and choose a feedback resistor. For this example we will use $10K\Omega$.



2. Calculate R1:
   Imagine the binary input is set to 0001_2, the desired output will be $-1V$.
   (a) Find V_{R5}: Because the V+ Pin3 is at 0V, V- Pin 2 will also be at 0V (with feedback). This means that $V_{R5} = 1V$ with polarities left to right equal to + -.
   (b) Calculate I_{R5}:
      - $I_{R5} = \frac{V_{R5}}{R5} = \frac{1V}{10K\Omega} = 100\mu A$
   (c) Determine V1: A TTL high will be approximately 5V.
   (d) Determine V_{R1}: $V_{R1} = V1 = 5V$
   (e) Determine I_{R1} : $I_{R1} = I_{R5} = 100\mu A$
   (f) Calculate $R1$: $R1 = \frac{V_{R1}}{I_{R1}} = \frac{5V}{100\mu A} = 50K\Omega$

3. Calculate R2:
   Imagine the binary input is set to 0010_2, the desired output will be $-2V$.
   (a) Find V_{R5}: Because the V+ Pin3 is at 0V, V- Pin 2 will also be at 0V (with feedback). This means that $V_{R5} = 2V$ with polarities left to right equal to + -.
   (b) Calculate I_{R5}:
      - $I_{R5} = \frac{V_{R5}}{R5} = \frac{2V}{10K\Omega} = 200\mu A$
   (c) Determine V2: A TTL high will be approximately 5V.
   (d) Determine V_{R2}: $V_{R2} = V2 = 5V$
   (e) Determine I_{R2} : $I_{R2} = I_{R5} = 200\mu A$
   (f) Calculate $R2$: $R2 = \frac{V_{R2}}{I_{R2}} = \frac{5V}{200\mu A} = 25K\Omega$

4. Calculate R3:
   Imagine the binary input is set to 0100_2, the desired output will be $-4V$.

(a) Find V_{R5}: Because the V+ Pin3 is at 0V, V- Pin 2 will also be at 0V (with feedback). This means that $V_{R5} = 4V$ with polarities left to right equal to + -.

(b) Calculate I_{R5}:

- $I_{R5} = \frac{V_{R5}}{R5} = \frac{4V}{10K\Omega} = 400\mu A$

(c) Determine V3: A TTL high will be approximately 5V.

(d) Determine V_{R3}: $V_{R3} = V3 = 5V$

(e) Determine I_{R3} : $I_{R3} = I_{R5} = 400\mu A$

(f) Calculate $R3$: $R3 = \frac{V_{R3}}{I_{R3}} = \frac{5V}{400\mu A} = 12.5K\Omega$

5. Calculate R4:
Imagine the binary input is set to 1000_2, the desired output will be $-8V$.

(a) Find V_{R5}: Because the V+ Pin3 is at 0V, V- Pin 2 will also be at 0V (with feedback). This means that $V_{R5} = 8V$ with polarities left to right equal to + -.

(b) Calculate I_{R5}:

- $I_{R5} = \frac{V_{R5}}{R5} = \frac{8V}{10K\Omega} = 800\mu A$

(c) Determine V4: A TTL high will be approximately 5V.

(d) Determine V_{R4}: $V_{R4} = V4 = 5V$

(e) Determine I_{R4} : $I_{R4} = I_{R5} = 800\mu A$

(f) Calculate $R4$: $R4 = \frac{V_{R4}}{I_{R4}} = \frac{5V}{800\mu A} = 6.25K\Omega$

6. Calculate R6:

(a) $R6 = \frac{1}{\frac{1}{R1}+\frac{1}{R2}+\frac{1}{R3}+\frac{1}{R4}+\frac{1}{R5}} = \frac{1}{\frac{1}{50K\Omega}+\frac{1}{25K\Omega}+\frac{1}{12.5K\Omega}+\frac{1}{6.25K\Omega}+\frac{1}{10K\Omega}}$

(b) $R6 = 2.5K\Omega$

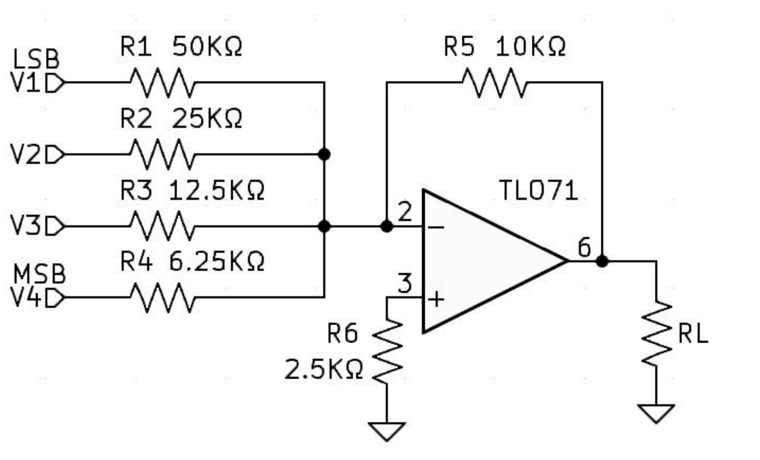


Figure 8.2: Designed Binary Weighted Digital to Analog Converter

8.3.3 R-2R DAC

R-2R Digital-to-Analogue Converter, or DAC, is a data converter which use two precision resistor to convert a digital binary number into an analogue output signal proportional to the value of the digital number - Electronics Tutorials [17]

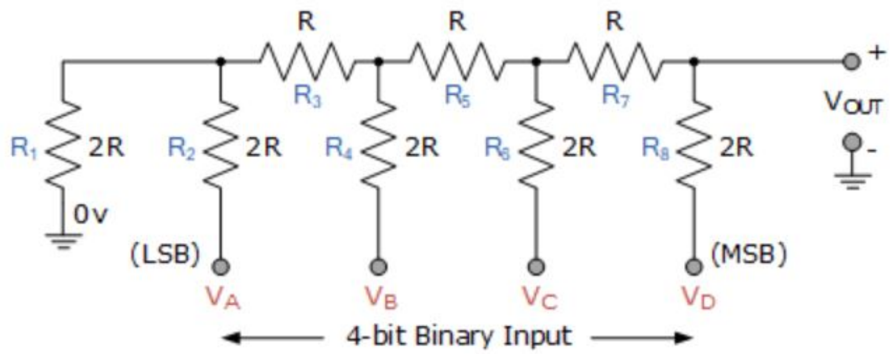


Figure 8.3: R-2R Digital to Analog Converter [17]

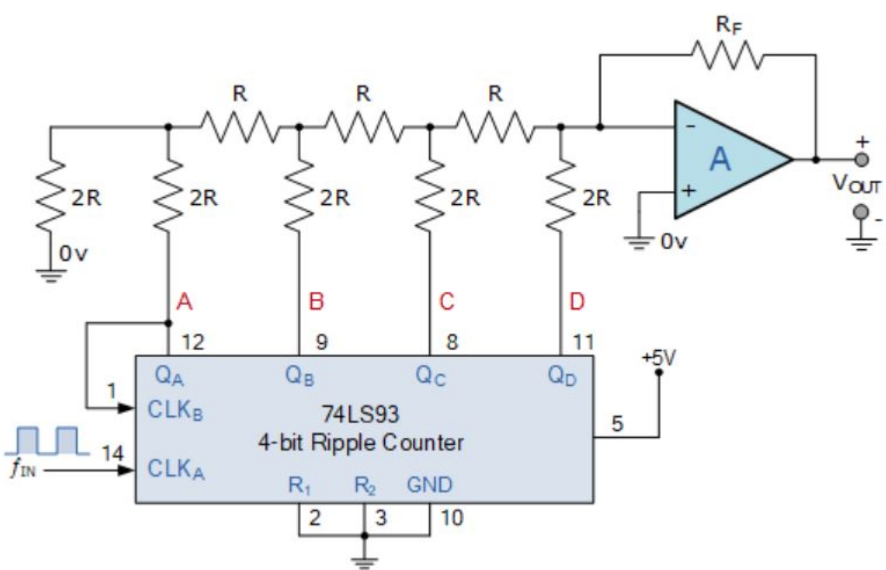


Figure 8.4: R-2R Digital to Analog Converter with 4-Bit Counter [17]

8.3.4 DAC0808

The DAC0808 is an 8-bit monolithic digital-to-analog converter (DAC) featuring a full scale output current settling time of 150 ns while dissipating only 33 mW with ±5V supplies. No reference current (IREF) trimming is required for most applications since the full scale output current is typically ±1 LSB of 255 IREF/256. Relative accuracies of better than ±0.19% assure 8-bit monotonicity and linearity while zero level output current of less than 4 µA provides 8-bit zero accuracy for IREF≥2 mA. The power supply currents of the DAC0808 is independent of bit codes, and exhibits essentially constant device characteristics over the entire supply voltage range. The DAC0808will interface directly with popular TTL, DTL or CMOS logic levels, and is a direct replacement for the MC1508/MC1408. For higher speed applications, see DAC0800 data sheet.
- DAC0808 Data Sheet [17]

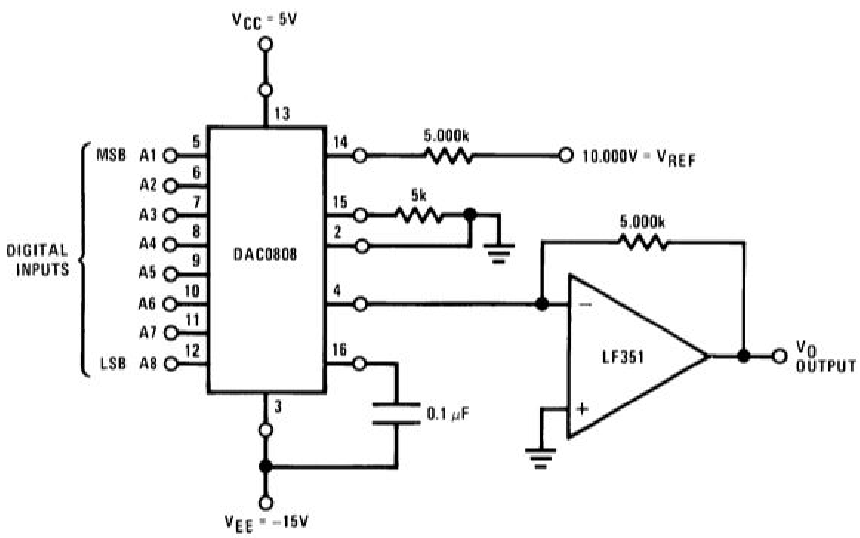


Figure 8.5: DAC0808 Typical Application [18]

8.4 Analog to Digital Conversion

8.4.1 Flash ADC (Direct Conversion ADC)

- **Principle:** Uses a bank of comparators, each one comparing the input signal to a unique reference voltage.
- **Advantages:** Very fast, as conversion happens in a single step.
- **Disadvantages:** Requires a large number of comparators, leading to high power consumption and complexity, especially for high resolutions.
- **Applications:** High-speed applications like digital oscilloscopes, radar, and high-speed data acquisition.

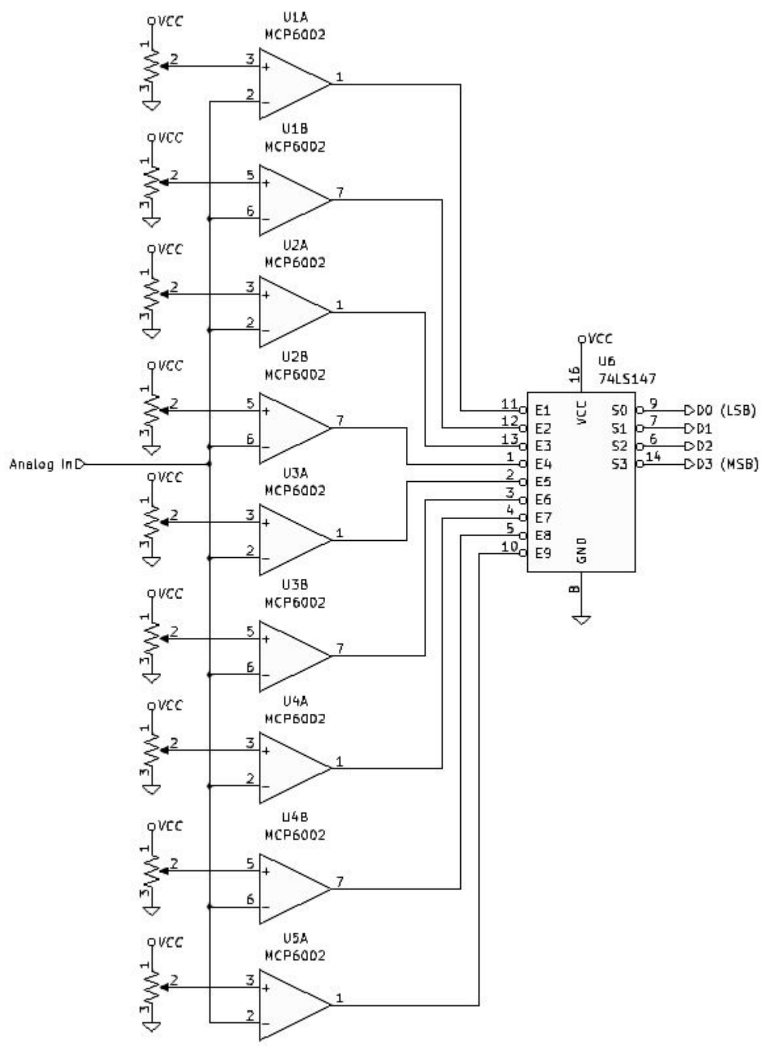


Figure 8.6: ADC Flash (Direct Conversion) 4-Bit

8.4.2 Additional Resources

- Flash ADC [19]
- Choosing the best ADC architecture [15]
- SAR and Delta-Sigma: Basic operation [20]
- Pipeline ADCs Come of Age [21]

8.5 DAC to ADC Circuit Integration

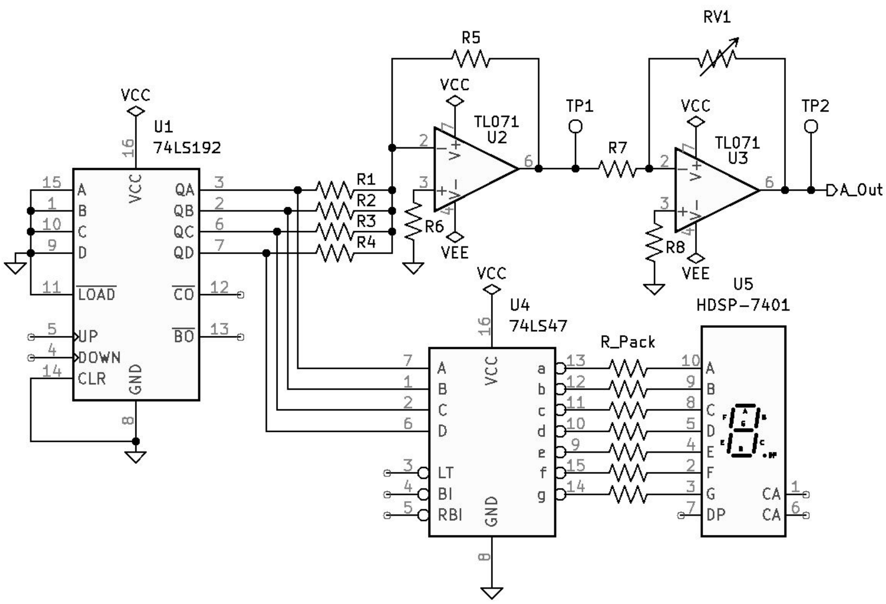


Figure 8.7: Design Test Circuit 1

Assignment:

1. Identify each IC used in Figure 8.7 Design Test Circuit 1 and include the following.
   - Part Description
   - Key Features
   - Key Applications
   - Important Specifications
2. Analyze the circuit and determine the overall function.
3. Verify that all wiring is correct.
4. Calculate the appropriate resistor values and the resistor current, voltage, and power.

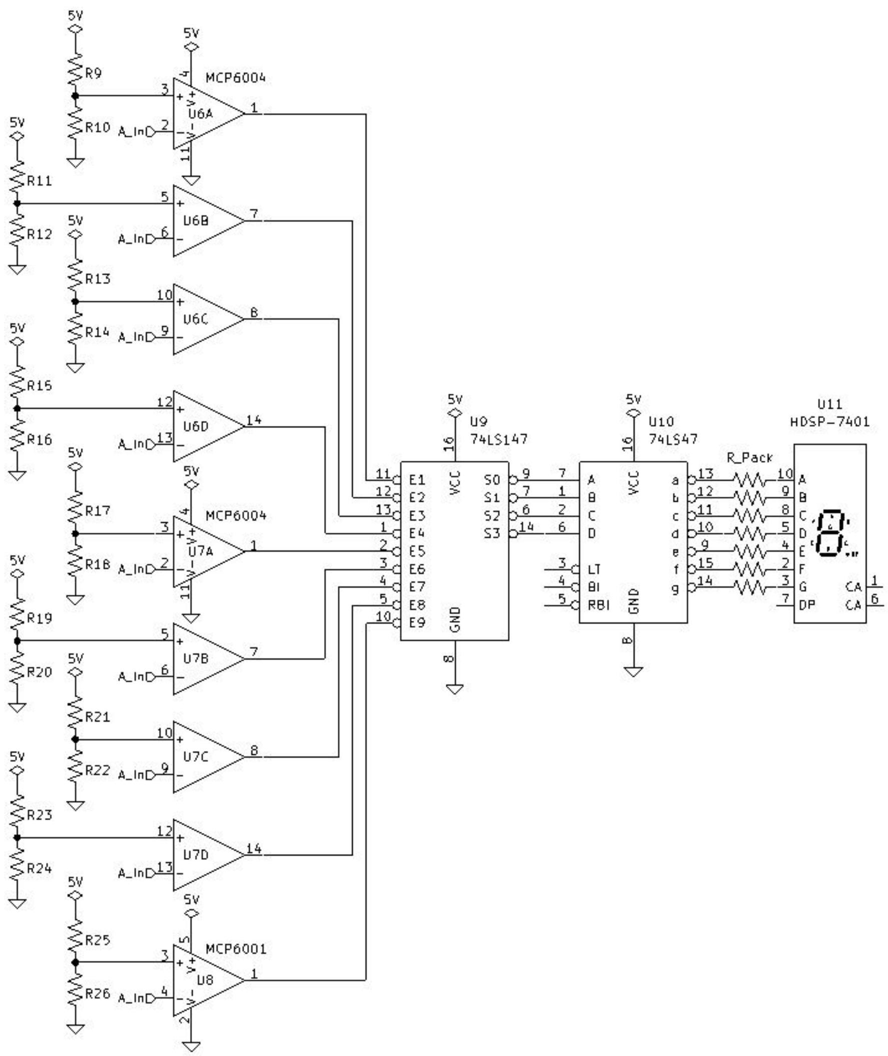


Figure 8.8: Design Test Circuit 2

Assignment:

1. Identify each IC used in Figure 8.8 Design Test Circuit 2 and include the following.
   - Part Description
   - Key Features
   - Key Applications
   - Important Specifications
2. Analyze the circuit and determine the overall function.
3. Verify that all wiring is correct.
4. Calculate the appropriate resistor values and the resistor current, voltage, and power.

Week 9

Linear Regulated Power Supply Project Part 1

Week 10

Linear Regulated Power Supply Project Part 2

Week 11a

Switching Diodes

11a.1 Objectives:

Understand the Role of Switching Diodes.

- Explain the purpose and importance of switching diodes in electronic circuits.
- Differentiate switching diodes from other types of diodes based on functionality and application.

Develop a Theoretical Foundation.

- Describe the basic operation of a PN junction and its relevance to switching diodes.
- Explain the concepts of forward and reverse biasing in the context of high-speed switching.

Identify Key Characteristics and Parameters.

- Analyze the I-V characteristics of switching diodes, including forward voltage drop and reverse leakage current.
- Interpret key parameters such as reverse recovery time, turn-on time, and maximum voltage/current ratings.
- Understand the impact of junction capacitance and charge storage on switching performance.

Apply Knowledge to Circuit Design

- Demonstrate the integration of switching diodes in basic and practical circuits such as clippers, clamps, and logic gates.
- Analyze and predict circuit behavior based on diode properties.

By the end of this module, students should have a solid theoretical understanding of switching diodes.

11a.2 General Information:

A diode is a fundamental electronic component that allows current to flow in one direction while blocking it in the opposite direction. Its primary purpose is to control the direction of electrical current in a circuit, making it essential for rectification, switching, signal demodulation, and protection. In rectifier circuits, diodes convert alternating current (AC) to direct current (DC), a vital process for powering electronic devices. Switching diodes enable high-speed transitions between conducting and non-conducting states, crucial for digital logic circuits and communication systems. Additionally, diodes protect sensitive components by blocking reverse currents or voltage spikes, ensuring the reliability and safety of electronic systems. Their versatility and functionality make diodes indispensable in a wide range of applications, from simple circuits to complex electronic systems.

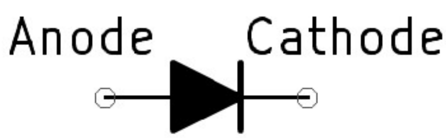


Figure 11a.1: Diode Schematic Symbol with Anode and Cathode label

11a.2.1 Diode Terminology

- **Diode** is a semiconductor device with a single PN junction that conducts current in only one direction.
- A **Forward-Biased** diode allows current to flow, exhibits a voltage drop of approximately 0.7V, and has a very low resistance. The anode must be positive relative to the cathode for the diode to be forward-biased.
- A **Reverse-Biased** diode behaves like an open switch, preventing current flow and exhibiting very high resistance. The anode must be negative relative to the cathode for the diode to be reverse-biased.
- A **Clipper or Limiter** is a diode circuit that clips off or removed part of a waveform above and/or below a specific voltage level.
- A **Clamper** is a circuit that adds a DC level to an AC waveform using a diode and a capacitor.
- A **Rectifier** is a circuit that converts alternating current (AC) into direct current (DC), typically using diodes.
- A **Rectifying Diode** is a semiconductor device primarily used to convert alternating current (AC) to direct current (DC) by allowing current to flow in only one direction.

- A **Switching Diode** is a semiconductor device designed to rapidly switch between conducting and non-conducting states, making it ideal for high-speed signal and digital circuit applications.
- A **Silicon Diode** is a semiconductor device made from silicon, commonly used for rectification and protecting circuits from reverse voltage, with a typical forward voltage drop of about 0.7V.
- A **Germanium Diode** is a semiconductor device made from germanium, known for its lower forward voltage drop (approximately 0.3V) and faster response time compared to silicon diodes, often used in low-voltage applications; while less common today, they are still manufactured for specialized uses such as vintage audio equipment, RF circuits, and precision low-voltage designs.
- A **Schottky Diode** is a semiconductor device characterized by its low forward voltage drop and fast switching speed, commonly used in high-frequency and power applications.

11a.3 Diode General Characteristics and Formulas:

- Forward bias voltage, V_F=0.7V (silicon) and 0.3V (germanium)
- Forward current, $I_F \approx$ 10mA up to Max forward current (data sheet)
- Diode power, $P_D = V_F \times I_F$
- Max forward current, $I_{F(Max)} = \frac{P_{D(Max)}}{V_F}$
- Reverse voltage, $V_R \approx 0\ to - 75V$
- Reverse current (I_R), " The reverse current I_R is at first equal to I_F: then it falls off to the reverse leakage current level." [2].
- Reverse leakage current, $I_S \approx 0.05\mu A$
- Reverse breakdown voltage, $VR_{Max} \approx 75V$
- Reverse recovery time, $t_{rr} \approx 4nS\ to\ 50nS$
  - The speed with which a diode can be switched is determined by the ***reverse recovery time*** of the device. [2, p. 76]
  - The ***reverse recovery time*** (t_{rr}) is the time required for the reverse current to fall to I_S. [2, p. 79]

11a.4 Diode Static Resistance

Diode Static Resistance (R_D) refers to the resistance offered by a diode under a steady-state condition, either in forward or reverse bias. It is calculated as the ratio of the voltage across the diode to the current flowing through it the diode:

Diode Forward Biased Static Resistance

Imagine a forward-biased silicone diode one that has a typical forward voltage of 0.7 volts and a forward current of $10mA$. Use Ohm's Law to calculate the forward biased static resistance of the diode.

$$R_D = \frac{V_F}{I_F}$$

- $R_D = \frac{0.7V}{10mA}$
- $R_D = 70\Omega$ (relatively small static resistance when forward biased, acting like a closed switch, notice the resistance will go down as current is increased.)

Diode Reverse Biased Static Resistance

Imagine a reverse-biased silicone diode, one that has a reverse voltage of 50 volts and a reverse leakage current of $0.5\mu A$. Use Ohm's Law to calculate the reverse biased static resistance of the diode.

$$R_D = \frac{V_R}{I_S}$$

- $R_D = \frac{50V}{0.05\mu A}$
- $R_D = 1G\Omega$ (large static resistance when reverse biased, acting like an open switch)

11a.5 Diode Dynamic Resistance

Similar to the $r'e$ of a bipolar junction transistor, when AC voltage is applied to a diode, it exhibits a dynamic resistance $r'd$ which can be calculated using the following formula.

$$r'd = \frac{26mV}{I_F}$$

11a.6 Reverse Recovery Time and Frequency Response

Practical design considerations for switching diodes

The diode's recovery time (t_{rr}) can significantly impact the critical high-frequency performance of a circuit. To minimize this effect, a diode with a recovery time of at least ten times faster than the desired rise or fall time should be used.

- $t_{rr} \leq \frac{Time_{Rise}}{10}$
- $Time_{Rise} \geq (t_{rr} \times 10)$
- $FC_{High} = \frac{0.35}{Time_{Rise}}$

11a.7 Data Sheet Information:

DEC

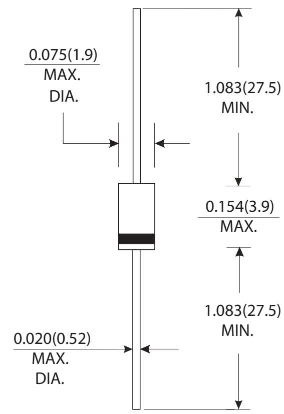
1N60, 1N60P

GERMANIUM DIODES

Features

- Metal silicon junction, majority carrier conduction
- High current capability, Low forward voltage drop
- Extremely low reverse current I_R
- Ultra speed switching characteristics
- Small temperature coefficient of forward characteristics
- Satisfactory Wave detection efficiency
- For use in RECORDER, TV, RADIO, TELEPHONE as detectors, super high speed switching circuits, small current rectifier

DO-35(GLASS)



Dimensions in inches and (millimeters)

Mechanical Data

- Case : DO-35 glass case
- Polarity : Color band denotes cathode end
- Weight : Approx. 0.13 gram

Absolute Ratings (Limiting Values)

Symbols	Parameters		Value 1N60	Value 1N60P	Units
V_{RRM}	Zenerepetitive Peak Reverse Voltage		40	45	Volts
I_F	Forward Continuous Crrent	T_A=25℃	30	50	mA
I_{FSM}	Peak Forward Surge Current(t=1S)		150	500	mA
T_{STG}/T_J	Storage junction Temperature Range		-65 to+125		℃
T_L	Maximum Lead Temperature for soldering 10S at 4mm from Case		230		℃

Electrical characteristics

Symbols	Parameters	Test Conditions		Value Min	Value Typ.	Value Max.	Units
V_F	Forward Voltage	I_F=1mA	1N60		0.32	0.5	Volts
			1N60P		0.24	0.5	
		I_F=30mA	1N60		0.65	1.0	
		I_F=200mA	1N60P		0.65	1.0	
I_R	Reverse Current	V_R=15V	1N60		0.1	0.5	µA
			1N60P		0.5	1.0	
C_J	Junction Capacitance	V_R=1V f=1MHz	1N60		2.0		pF
		V_R=10V f=1MHz	1N60P		6.0		
η	Detection Effcienc(See diagram 4)	V_I=3V f=30MHz C_L=10pF R_L=3.8kΩ			60		%
trr	Revese Recovery time	I_F=I_R=1mA Irr=1mA R_C=100Ω				1	ns
$R_{\Theta JA}$	Junction Amblent Thermal Resistance				400		℃/W

DEC

RATINGS AND CHARACTERISTIC CURVES 1N60P

FIG.1-FORWARD CURRENT VERSUS FORWARD VALTAGE(TYPICAL VALUES)

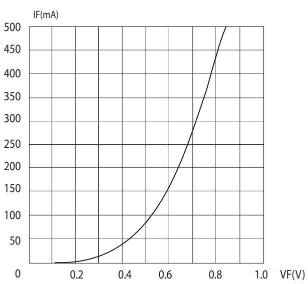


FIG.2-REVERSE CURRENT VERSUS CONTINUOUS REVERSE VOLTAGE

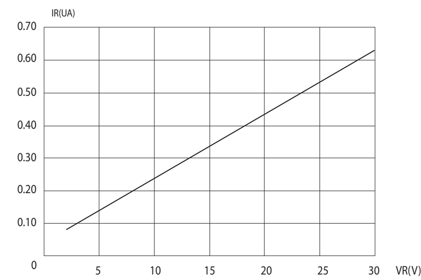


FIG.3-JUNCTION CAPACITANCE VERSUS CONTINUOUS REVERSE APPLIED VOLTAGE

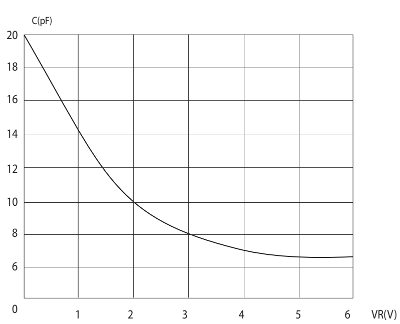
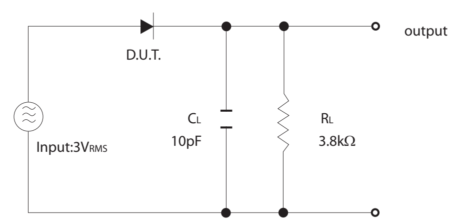


FIG.4-DETECTION EFFICIENCY MEASUREMENT CIRCUIT



VISHAY.

www.vishay.com

1N4148

Vishay Semiconductors

Small Signal Fast Switching Diodes

FEATURES

- Silicon epitaxial planar diode
- Electrically equivalent diode:1N914
- Material categorization: for definitions of compliance please see www.vishay.com/doc?99912

Pb

e3

RoHS COMPLIANT

HALOGEN FREE

APPLICATIONS

- Extreme fast switches

LINKS TO ADDITIONAL RESOURCES

3D Models | Marking | Parametric Search | Order Samples

MECHANICAL DATA

Case: DO-35 (DO-204AH)

Weight: approx. 105 mg

Cathode band color: black

Packaging codes / options:

TR/10K per 14" reel (52 mm tape), 50K/box

TAP/10K per ammopack (52 mm tape), 50K/box

PARTS TABLE

PART	ORDERING CODE	TYPE MARKING	CIRCUIT CONFIGURATION	REMARKS
1N4148	1N4148-TAP or 1N4148TR	V4148	Single	Tape and reel / ammopack

ABSOLUTE MAXIMUM RATINGS (T_{amb} = 25 °C, unless otherwise specified)

PARAMETER	TEST CONDITION	SYMBOL	VALUE	UNIT
Repetitive peak reverse voltage		V_{RRM}	100	V
Reverse voltage		V_R	75	V
Peak forward surge current	t_p = 1 µs	I_{FSM}	2	A
Repetitive peak forward current		I_{FRM}	500	mA
Forward continuous current		I_F	300	mA
Average forward current	$V_R = 0$	$I_{F(AV)}$	150	mA
Power dissipation	l = 4 mm, T_L = 45 °C	P_{tot}	440	mW
	l = 4 mm, $T_L \leq$ 25 °C	P_{tot}	500	mW

THERMAL CHARACTERISTICS (T_{amb} = 25 °C, unless otherwise specified)

PARAMETER	TEST CONDITION	SYMBOL	VALUE	UNIT
Thermal resistance junction to ambient air	l = 4 mm, T_L = constant	R_{thJA}	350	K/W
Junction temperature		T_j	175	°C
Storage temperature range		T_{stg}	-65 to +150	°C

ELECTRICAL CHARACTERISTICS (T_{amb} = 25 °C, unless otherwise specified)						
PARAMETER	**TEST CONDITION**	**SYMBOL**	**MIN.**	**TYP.**	**MAX.**	**UNIT**
Forward voltage	I_F = 10 mA	V_F			1	V
Reverse current	V_R = 20 V	I_R			25	nA
	V_R = 20 V, T_j = 150 °C	I_R			50	μA
	V_R = 75 V	I_R			5	μA
Breakdown voltage	I_R = 100 μA, t_p/T = 0.01, t_p = 0.3 ms	$V_{(BR)}$	100			V
Diode capacitance	V_R = 0 V, f = 1 MHz, V_{HF} = 50 mV	C_D			4	pF
Rectification efficiency	V_{HF} = 2 V, f = 100 MHz	η_r	45			%
Reverse recovery time	$I_F = I_R$ = 10 mA, i_R = 1 mA	t_{rr}			8	ns
	I_F = 10 mA, V_R = 6 V, i_R = 0.1 x I_R, R_L = 100 Ω	t_{rr}			4	ns

TYPICAL CHARACTERISTICS (T_{amb} = 25 °C, unless otherwise specified)

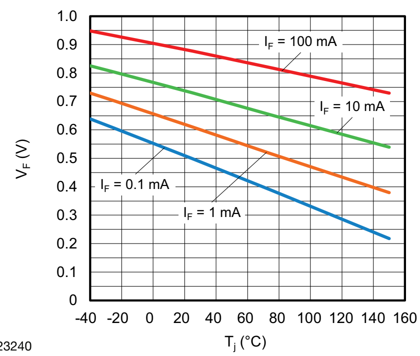


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Fig. 1 - Typical Forward Voltage vs. Junction Temperature

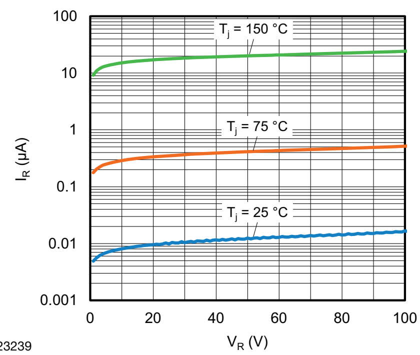

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Fig. 3 - Typical Reverse Leakage Current vs. Reverse Voltage

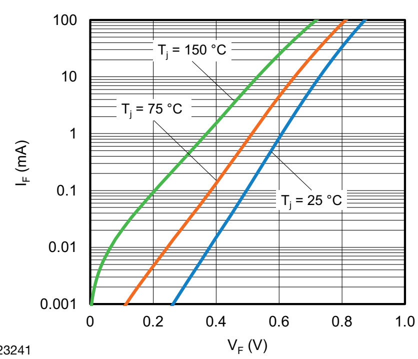

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Fig. 2 - Forward Current vs. Forward Voltage

FAIRCHILD

November 2014

1N4001 - 1N4007
General-Purpose Rectifiers

Features

- Low Forward Voltage Drop
- High Surge Current Capability

DO-41
COLOR BAND DENOTES CATHODE

Ordering Information

Part Number	Top Mark	Package	Packing Method
1N4001	1N4001	DO-204AL (DO-41)	Tape and Reel
1N4002	1N4002	DO-204AL (DO-41)	Tape and Reel
1N4003	1N4003	DO-204AL (DO-41)	Tape and Reel
1N4004	1N4004	DO-204AL (DO-41)	Tape and Reel
1N4005	1N4005	DO-204AL (DO-41)	Tape and Reel
1N4006	1N4006	DO-204AL (DO-41)	Tape and Reel
1N4007	1N4007	DO-204AL (DO-41)	Tape and Reel

Absolute Maximum Ratings

Stresses exceeding the absolute maximum ratings may damage the device. The device may not function or be operable above the recommended operating conditions and stressing the parts to these levels is not recommended. In addition, extended exposure to stresses above the recommended operating conditions may affect device reliability. The absolute maximum ratings are stress ratings only. Values are at T_A = 25°C unless otherwise noted.

Symbol	Parameter	Value							Unit
		1N 4001	1N 4002	1N 4003	1N 4004	1N 4005	1N 4006	1N 4007	
V_{RRM}	Peak Repetitive Reverse Voltage	50	100	200	400	600	800	1000	V
$I_{F(AV)}$	Average Rectified Forward Current .375 " Lead Length at T_A = 75°C	1.0							A
I_{FSM}	Non-Repetitive Peak Forward Surge Current 8.3 ms Single Half-Sine-Wave	30							A
I^2t	Rating for Fusing (t < 8.3 ms)	3.7							A^2sec
T_{STG}	Storage Temperature Range	-55 to +175							°C
T_J	Operating Junction Temperature	-55 to +175							°C

Thermal Characteristics

Values are at T_A = 25°C unless otherwise noted.

Symbol	Parameter	Value	Unit
P_D	Power Dissipation	3.0	W
$R_{\theta JA}$	Thermal Resistance, Junction-to-Ambient	50	°C/W

Electrical Characteristics

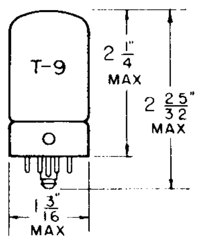
Values are at T_A = 25°C unless otherwise noted.

Symbol	Parameter	Conditions	Value	Unit
V_F	Forward Voltage	I_F = 1.0 A	1.1	V
I_{rr}	Maximum Full Load Reverse Current, Full Cycle	T_A = 75°C	30	µA
I_R	Reverse Current at Rated V_R	T_A = 25°C	5.0	µA
		T_A = 100°C	50	
C_T	Total Capacitance	V_R = 4.0 V, f = 1.0 MHz	15	pF

TENTATIVE DATA 7Y4

TUNG-SOL

DOUBLE DIODE

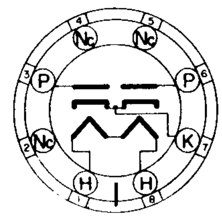


GLASS BULB

UNIPOTENTIAL CATHODE

HEATER
6.3 VOLTS 500 MA.
AC OR DC

ANY MOUNTING POSITION



BOTTOM VIEW
LOCK-IN 8 PIN BASE

THE 7Y4 IS A HEATER TYPE HIGH VACUUM TWIN DIODE USING THE LOCK-IN CONSTRUCTION. IT IS INTENDED FOR USE AS A FULL-WAVE RECTIFIER IN EITHER AC OR STORAGE BATTERY OPERATED EQUIPMENT WHERE ECONOMY OF HEATER POWER IS DESIRED.

RATINGS

INTERPRETED ACCORDING TO RMA STANDARD M8-210

HEATER VOLTAGE	6.3	VOLTS
MAXIMUM DC HEATER-CATHODE VOLTAGE	450	VOLTS
MAXIMUM PEAK INVERSE VOLTAGE	1 250	VOLTS
MAXIMUM AC PLATE VOLTAGE (RMS)CONDENSER INPUT	325	VOLTS
MAXIMUM AC PLATE VOLTAGE (RMS) CHOKE INPUT	450	VOLTS
MAXIMUM STEADY STATE PEAK PLATE CURRENT EACH PLATE	210	MA.
MAXIMUM OUTPUT CURRENT	70	MA.
TUBE VOLTAGE DROP (MEASURED WITH TUBE CONDUCTING 70 MA. EACH PLATE)	22	VOLTS

TYPICAL OPERATING CONDITIONS AND CHARACTERISTICS

FULL WAVE RECTIFIER

CONDENSER INPUT TO FILTER

HEATER VOLTAGE	6.3	VOLTS
HEATER CURRENT	500	MA.
AC PLATE VOLTAGE EACH PLATE (RMS)	325	VOLTS
DC OUTPUT CURRENT	70	MA.
MINIMUM PLATE SUPPLY IMPEDANCE EACH PLATE[A]	150	OHMS

[A] WHEN A FILTER CONDENSER LARGER THAN 40 μF IS USED, IT MAY BE NECESSARY TO INCREASE THE SPECIFIED PLATE SUPPLY IMPEDANCE.

CHOKE INPUT TO FILTER

HEATER VOLTAGE	6.3	VOLTS
HEATER CURRENT	500	MA.
AC PLATE VOLTAGE EACH PLATE (RMS)	450	VOLTS
DC OUTPUT CURRENT	70	MA.
MINIMUM VALUE OF INPUT CHOKE	10	HENRYS

PRINTED IN U. S. A.

PLATE 2107
NOV. 1, 1948

SIMILAR TYPE REFERENCE: Ratings and characteristics somewhat similar to types 6X5GT and 84.

Week 11b

Switching Transistors

11b.1 Objectives

1. **Theory of Operation:**
   - Describe how a transistor operates in different regions (cutoff, active, and saturation).
   - Explain the concept of a transistor as a switch.
2. **Switching Characteristics:**
   - Analyze the switching behavior of transistors, including turn-on and turn-off times.
   - Understand parameters such as rise-time, fall-time, delay-time, and storage-time.
3. **Circuit Design and Analysis:**
   - Design simple switching circuits using transistors.
   - Analyze and calculate the required component values for desired switching performance.
   - Understand the term Overdrive in terms of Switching Transistors and its effect on Turn-On and Turn-Off times.
   - Design an improved Switching Transistor circuit using a Commutating Capacitor.

11b.2 Switching Transistors Identification

- The switching transistor circuit is NOT AN AMPLIFIER!! It is easily identified because it is missing a biasing circuit: No Base Bias, No Emitter Bias, No Universal Bias, No Collector Feedback Bias...

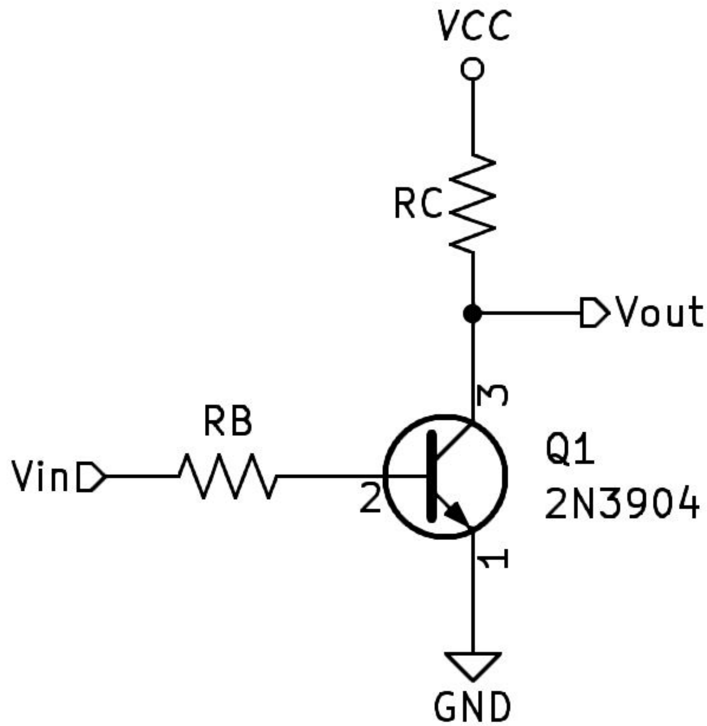


Figure 11b.1: Switching Transistor Circuit

- The switching transistor circuit is designed to operate, not in the Active Region, but in Saturation and Cutoff.
- Switching transistor circuits operate much more efficiently than amplifier circuits because they operate in saturation and cutoff and not the active region.
  - In saturation, IC is max and VCE is theoretically 0V. $P_Q = 0w$
  - In cutoff, IC is zero and VCE is max. $P_Q = 0w$
- The Common Emitter switch configuration is an inverting switch.
  - A high V_{in} will equal low V_{out}. The transistor is in saturation.
  - A low V_{in} will equal high V_{out}. The transistor is in cutoff.

11b.3 Switching Transistor Turn-On and Turn-Off Times

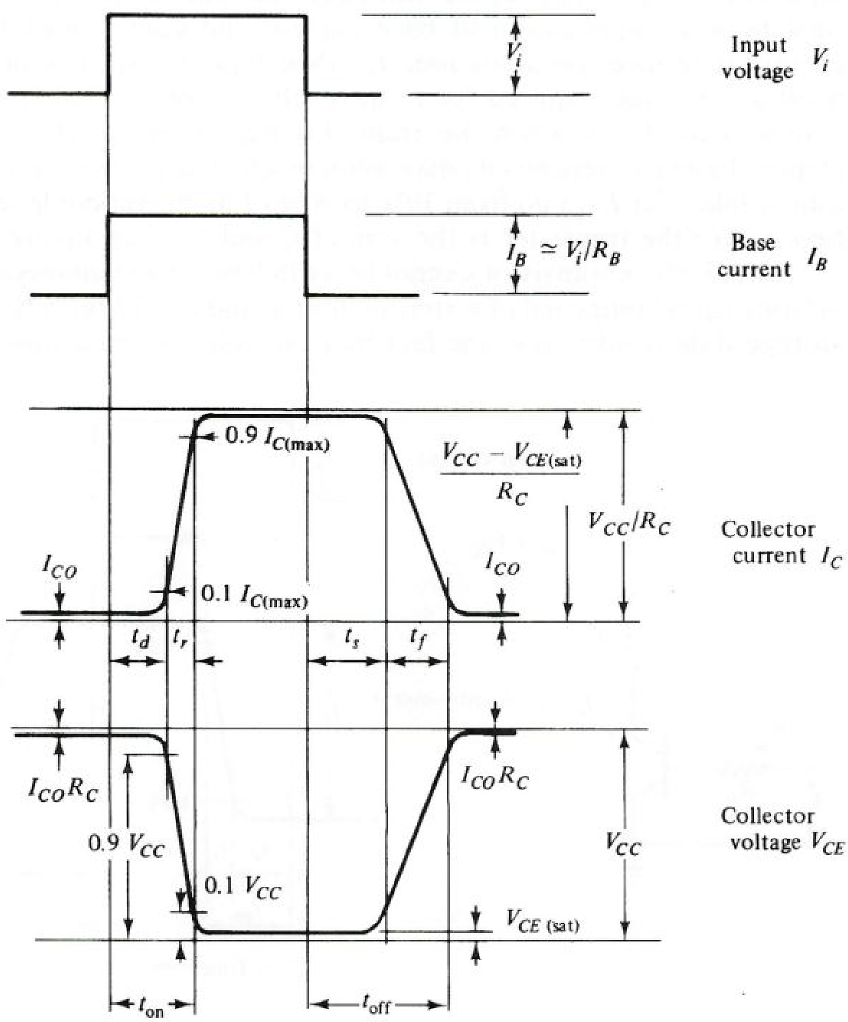


Figure 11b.2: Switching Transistor Turn-On and Turn-Off Characteristics - Bell [2]

- t_d = Delay Time
- t_r = Rise Time. The transistor is turning on. Rise Time occurs when the input signal is transitioning from 10% to 90% of the signal voltage. The collector current will begin to flow from its 10% to 90% of IC_{sat} during the Rise Time.
- Turn-On Time = $t_d + t_r$
- t_f = Fall Time
- t_s = Storage Time

- Turn-Off Time $= t_s + t_f$

11b.3.1 2N3904 Turn-On & Turn-Off Times

Switching Characteristics						
Delay Time	t_d	V_{CC} = 3V, V_{EB} = 0.5V, I_C = 10mA, I_{B1} = 1mA	–	–	35	ns
Rise Time	t_r		–	–	35	ns
Storage Time	t_s	V_{CC} = 3V, I_C = 10mA, I_{B1} = I_{B2} = 1mA	–	–	200	ns
Fall Time	t_f		–	–	50	ns

Figure 11b.3: 2N3904 Switching Characteristics - NTE [22]

- 2N3904 Turn-On Time $= t_d + t_r$
  - 2N3904 Turn-On Time $= 35nS + 35nS$
  - 2N3904 Turn-On Time $= 70nS$
- 2N3904 Turn-Off Time $= t_s + t_f$
  - 2N3904 Turn-Off Time $= 200nS + 50nS$
  - 2N3904 Turn-Off Time $= 250nS$

11b.4 Switching Transistor Design

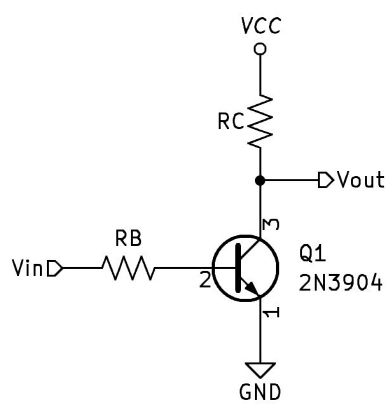


Figure 11b.4: Switching Transistor Circuit

Calculation Steps:

- $IC_{Sat} = \frac{VCC}{RC}$
- $IB_{Sat} = \frac{IC_{Sat}}{Beta_{Q1}}$
- $RB_{Sat} = \frac{V_{in} - Vbe_{Q1}}{IB_{Sat}}$

11b.5 Overdriving a Switching Transistor

Overdriving is achieved by increasing the switching transistors base current beyond IB_{Sat}. Overdriving the transistor switch will improve or decrease the Turn-On Time. However, the disadvantage is Overdriving will increase the Storage-Time which will increase the Turn-Off Time.

Overdrive by percentage

- To Overdrive by 10%, simply reduce RB_{Sat} by 10%. $RB_{10\%OD} = RB_{Sat} \times 0.9$
- To Overdrive by 20%, simply reduce RB_{Sat} by 20%. $RB_{20\%OD} = RB_{Sat} \times 0.8$
- Measure Turn-On and Turn-Off Times as Overdrive is being applied to the switching transistor. Repeat... until a decrease in RB no longer lowers the Turn-On Time.

11b.6 Improved Turn-On and Turn-Off Times using a Commutating Capacitor

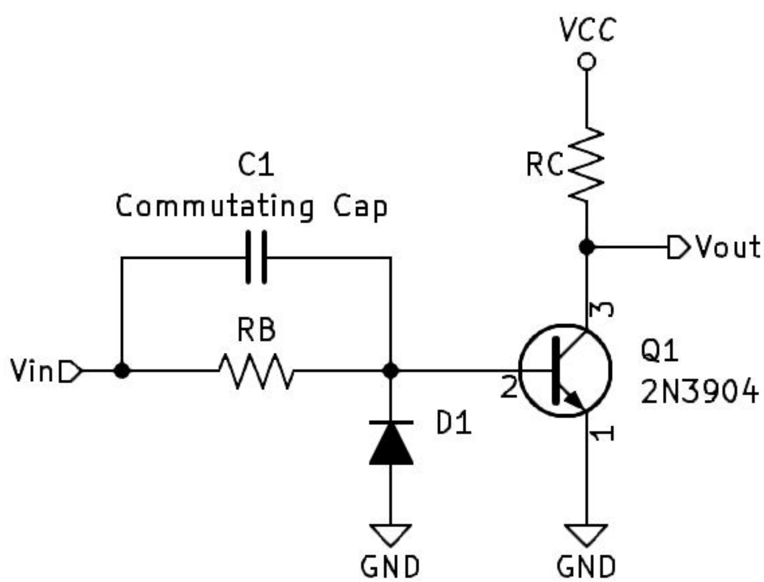


Figure 11b.5: Improved Switching Transistor Circuit

Adding a Commutating Capacitor will improve both Turn-On and Turn-Off Times for the switching transistor circuit.

11b.6.1 Commutating Capacitor Theory of Operation

- When V_{in} transitions from a low to a high, the Commutating Capacitor will initially act like a short causing the transistor to be overdriven and improving the Turn-On Time. Once the capacitor is charged ($V_{C1} = V_{in} - Vbe_{Q1}$) the current path is through RB which is designed to keep the transistor in saturation and not in overdrive.
- When V_{in} transitions from a high to a low, (assuming there was previously enough time to fully charge the capacitor) the voltage across the commutating capacitor and RB network is $V_{C1} = V_{in} - Vbe_{Q1}$ meaning that as V_{in} goes to 0V, the base of the transistor will see a negative voltage with respect to ground and its own emitter. This negative voltage at the base of the transistor will help turn off the transistor and because the transistor was previously in saturation and not in overdrive the Turn-Off Time is optimized.

Absolute Maximum Ratings:

Collector–Emitter Voltage, V_{CEO}	40V
Collector–Base Voltage, V_{CB}	60V
Emitter–Base Voltage, V_{EBO}	6V

Figure 11b.6: 2N3904 V_{EB} Max - NTE [22]

- Observe in Figure 11b.6 *2N3904 V_{EB} Max - NTE [22]* that the maximum negative voltage base to emitter is -6V ($Vmax_{EB} = 6V$, $Vmax_{BE} = -6V$). This means that if V_{in} is more that 6.7 volts the transistor could be damaged when V_{in} transitions to 0V due to the now negative voltage held by the capacitor. Diode D1 serves two functions, it is used to limit the negative voltage seen at V_{BE} to -0.7V; secondly, it removes RB from the discharge path of the Commutating Capacitor allowing for a faster discharge. With diode D1 forward biased, via the discharging Commutating Capacitor, the discharge path is now through the diode and R_{Gen} of the input generator. This allows for a rapid discharge of the Commutating Capacitor, improving the Recovery Time.

11b.6.2 Recovery Time and Maximum Frequency Calculations

1. Design a switching transistor circuit to operate in Saturation (do not overdrive).
2. Determine the Turn-On Time for the transistor the switching transistor used.
   - Turn-On Time = $t_d + t_r$
     - 2N3904 Turn-On Time = 35nS + 35nS
     - 2N3904 Turn-On Time = 70nS
3. Use the Turn-On Time to calculate the tau of the Commutating Capacitor.
   - $t_{on} = 0.1R_{Gen}C_C$
     - t_{on} =Turn-On Time

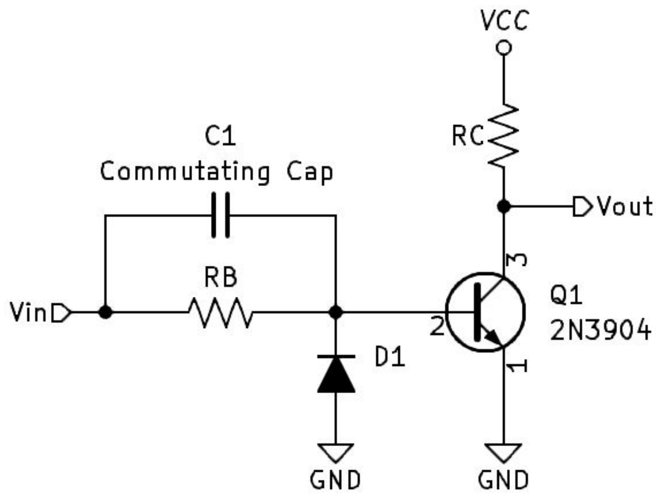
- making the tau ten times the turn-on time allows the capacitor to quickly charge while providing an initial overdrive current.
- R_{Gen} = Generator Resistance
- C_C = Commutating Capacitor

4. Inital Over-Drive Current Calculations:
   - $IB(max)_{Instantaneous} = \frac{V_{Gen} - V_{BE}}{R_{Gen}}$
5. Determine the Recovery Time t_{re}.
   - The Recovery Time is the time needed for the capacitor to charge or discharge 90% of its final voltage.
   - At 5tau the capacitor will be fully charged or fully discharged.
   - Solve for the tau at 90%:
     - $V_C = V_{fin} - (V_{fin} - V_{in})e^{\frac{-t_{re}}{RC}}$
     - $0.9 = 1 - (1 - 0)e^{\frac{-t_{re}}{RC}}$
     - $0.9 - 1 = -(1 - 0)e^{\frac{-t}{RC}}$
     - $-0.1 = -(1 - 0)e^{\frac{-t_{re}}{RC}}$
     - $-0.1 = -(1)e^{\frac{-t_{re}}{RC}}$
     - $\frac{-0.1}{-1} = e^{\frac{-t_{re}}{RC}}$
     - $0.1 = e^{\frac{-t_{re}}{RC}}$
     - $LN(0.1) = \frac{-t_{re}}{RC}$
     - $\frac{-t_{re}}{RC} = -2.303$
     - $-t_{re} = -2.303(RC)$
     - With diode D1 in the circuit, R_{Gen} becomes the primary resistance in the discharge path of the Commutating Capacitor, you could add the dynamic resistance of the diode.

   ✓ $t_{re} = 2.303(RC)$
6. Determine the maximum square wave input frequency f_{MaxCC} for the Commutating Capacitor Switching Transistor circuit:
   - $f_{MaxCC} \approx \frac{1}{2t_{re}}$

11b.7 Practice Questions:

1. Design a Switching Circuit using a Commutating Capacitor to improve both Turn-On and Turn-Off Times.



Given: $VCC = 9V$, $IC = 10mA$, $Beta_{min} = 50$, $V_{in} = 0\ to\ 5V$, $T_{On} = 50nS$, $R_{Gen} = 50\Omega$.

Find: RC, RB, & $C1$.

Solve:

$RC =$ ____________

$RB =$ ____________

$C1 =$ ____________

11b.8 Answers:

1. $RC = \underline{880\Omega}$

   $RB = \underline{21.45K\Omega}$

   $C1 = \underline{0.01\mu F}$

Week 11c

Multivibrators

11c.1 Objectives:

1. **Introduction to Multivibrators:**
   - Define what multivibrators are and their role in electronic circuits.
   - Differentiate between the types of multivibrators: astable, monostable, and bistable.
2. **Theory of Operation:**
   - Explain the basic principles and operation of each type of multivibrator.
   - Discuss the conditions under which each type operates and their typical applications.
3. **Astable Multivibrators:**
   - Describe the structure and function of an astable multivibrator.
   - Analyze the waveform outputs and timing characteristics.
   - Design and calculate component values for a given frequency and duty cycle.
4. **Monostable Multivibrators**
   - Explain the operation of a monostable multivibrator.
   - Discuss its use as a pulse generator.
   - Design circuits to produce a specific pulse width.
5. **Circuit Design and Analysis:**
   - Build and analyze multivibrator circuits using both discrete components and integrated circuits (ICs).

11c.2 Introduction:

A multivibrator is a versatile electronic circuit used to implement simple yet essential functionalities in various applications. Comprising resistors, capacitors, and transistors or integrated circuits, multivibrators are categorized into three types: astable, monostable, and bistable. Each type serves a distinct purpose: astable multivibrators generate continuous oscillations without requiring an external trigger, monostable multivibrators produce a single output pulse in response to an input trigger, and bistable multivibrators, also known as flip-flops, toggle between two stable states based on input signals. These circuits play a crucial role in timing, waveform generation, and digital logic operations, making them foundational elements in the design of clocks, pulse generators, memory storage devices, and other critical electronic systems.

11c.3 Schmitt Triggered Astable Multivibrator:

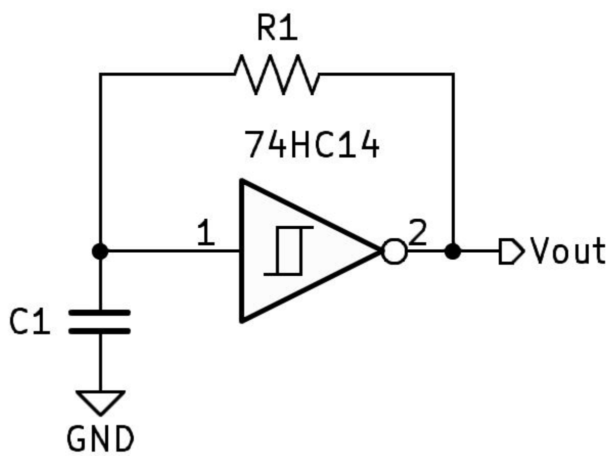


Figure 11c.1: Schmitt Triggered Astable Multivibrator

11c.3.1 SN74HC14 Specifications

- $Vout_H \approx 5V$
- $Vout_L \approx 0V$
- $V_{T+} \approx 2.5V$
- $V_{T-} \approx 1.6V$
- To get more accuracy use measured $Vout_H$, $Vout_L$, V_{T+}, and V_{T-}.

11c.3.2 Pulse Width Calculations:

- $V_C = V_{fin} - (V_{fin} - V_{in})e^{\frac{-t}{RC}}$
  - $V_{T+} = Vout_H - (Vout_H - V_{T-})e^{\frac{-PW}{RC}}$
  - $2.5V = 5V - (5V - 1.6V)e^{\frac{-PW}{RC}}$
  - $2.5V - 5V = -(5V - 1.6V)e^{\frac{-PW}{RC}}$
  - $-2.5V = -(3.4V)e^{\frac{-PW}{RC}}$
  - $\frac{2.5}{3.4} = e^{\frac{-PW}{RC}}$
  - $LN(\frac{2.5}{3.4}) = \frac{-PW}{RC}$
  - $-0.30748 = \frac{-PW}{RC}$
  - ✓ $PW = RC \times 0.30748$
  - The R value: The gate needs to have enough current Iin_H to allow the inverter to switch states. Generally speaking, we would want the capacitor current to be ten times larger that Iin_H at the time of the transition. This means that we may be instantaneously (for a short amount of time) exceeding the $Iout_H$ specification. Because of this we need to keep the R1 value small enough to maintain oscillations but large enough to protect the output current specifications/limitations of the inverter. If the R1 value is too big the circuit will not oscillate.
  - $R1 \approx \frac{Vout_H - V_{T+}}{2 \times Iout_H}$
    - $R1 \approx \frac{5V - 2.5V}{2 \times 400 \mu A}$
    - $R1 \approx \frac{2.5V}{800 \mu A}$

    ✓ $R1 \approx 3.125K\Omega$
  - $C \approx \frac{PW}{R1 \times 0.30748}$

    ✓ $C \approx \frac{PW}{3.125K\Omega \times 0.30748}$

11c.3.3 Pulse Space Calculations:

- $V_C = V_{fin} - (V_{fin} - V_{in})e^{\frac{-t}{RC}}$
  - $V_{T-} = Vout_L - (Vout_L - V_{T+})e^{\frac{-PS}{RC}}$
  - $1.6V = 0 - (0 - 2.5V)e^{\frac{-PS}{RC}}$
  - $1.6V = -(-2.5V)e^{\frac{-PS}{RC}}$
  - $1.6V = (2.5V)e^{\frac{-PS}{RC}}$
  - $\frac{1.6V}{2.5V} = e^{\frac{-PS}{RC}}$

- $\circ$ $LN(\frac{1.6V}{2.5V}) = \frac{-PS}{RC}$
- $\circ$ $-0.446287 = \frac{-PS}{RC}$
- $\circ$ $-0.446287 \times RC = -PS$
- ✓ $PS \approx 0.446287(3.125K\Omega \times C)$

- Build and measure, adjust the capacitor value to achieve the desired frequency.

11c.4 BJT Astable Multivibrator:

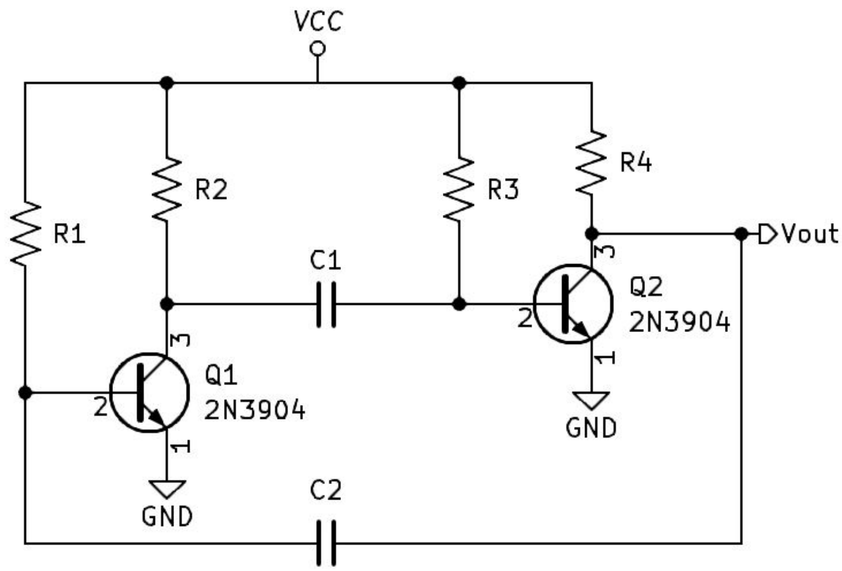


Figure 11c.2: Bipolar Junction Transistor Astable Multivibrator

11c.4.1 Introduction

The Bipolar Junction Transistor (BJT) Astable Multivibrator is a fundamental electronic circuit that generates a continuous square wave output, making it an essential component in timing and waveform generation applications. Unlike monostable or bistable multivibrators, an astable multivibrator has no stable states; it constantly switches between its two unstable states, producing a periodic oscillation without the need for an external trigger. Utilizing BJTs, capacitors, and resistors, this circuit operates by alternately driving the transistors into saturation and cutoff regions, resulting in a consistent toggling action. BJT astable multivibrators are widely used in applications such as clock pulse generation, LED flashers, and tone generators, providing a simple yet effective solution for generating periodic signals in various electronic devices.

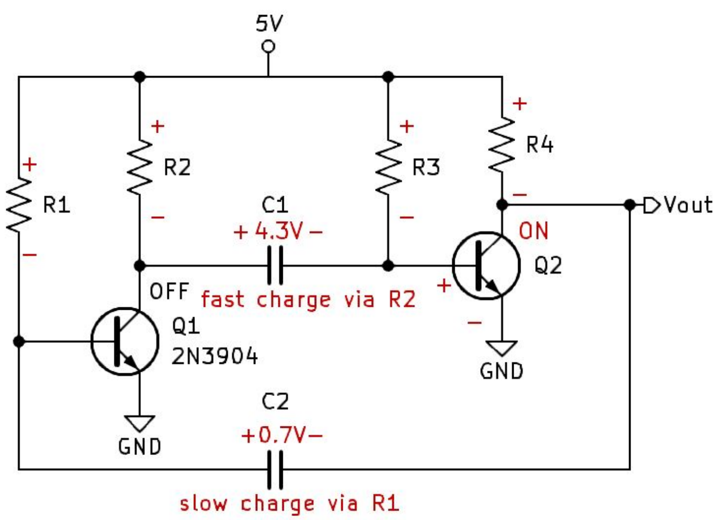
11c.4.2 Calculation Steps

Calculating R values:

1. The transistors should act like switches operating in the Saturation and Cuttoff Regions.
   (a) Choose the saturation current IC_{Sat}.
   (b) $R2\ \&\ R4 = \frac{VCC}{IC_{Sat}}$
   (c) $IB_{Sat} = \frac{IC_{Sat}}{Beta_{min}}$
   (d) $R1\ \&\ R3 = \frac{VCC - V_{BE}}{IB_{Sat}}$
   (e) Example: $VCC = 5V$, $IC_{Sat} = 10mA$, and $Beta_{Q1\&Q2} = 100$:
      i. $R2\ \&\ R4 = 500\Omega$
      ii. $R1\ \&\ R3 = 43K\Omega$

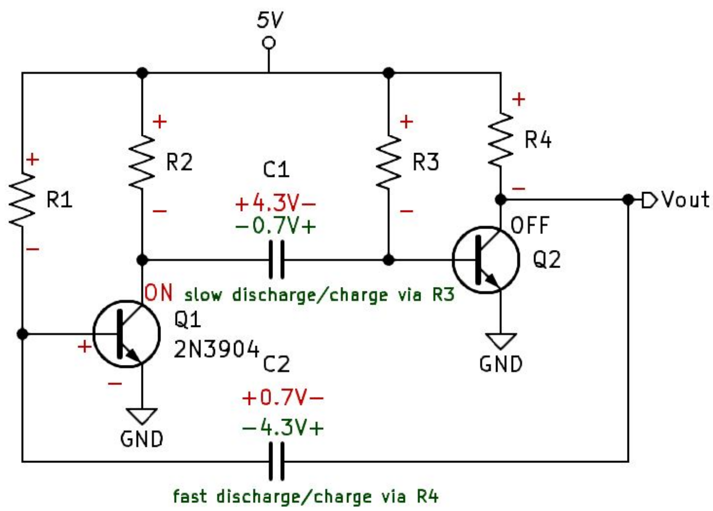
Calculating C values:

2. Notice R2 and R4 are relatively small compared to R1 and R3. Assuming C1 and C2 are equal and depending on which transistor is on, the capacitor with the charge path through R2 or R4 will change much faster than the other capacitor which will have either R1 or R3 to charge/discharge through.



Assuming that $VCC = 5V$ and Q2 is ON and Q1 is OFF.

(a) C1 (relative to C2) will charge quickly through the smaller RC resistor R2.
(b) C2 (relative to C1) will charge slower through the larger RB resistor R1.



(c) At the moment C2 reaches 0.7V, Q1 will turn on. When Q1 turns on it provides a path to ground, the capacitor C1 has been charged to 4.3V and now the base of Q2 has -4.3V VBE which will immediately force Q2 Off.

(d) Observe that the circuit will oscillate between these two states. Q1 and Q2 will oscillate between on and off opposite each other, and C1 constantly charge and discharge between 4.3V and -0.7V and C2 will charge and discharge between 0.7V and -4.3V. The timing of the circuit is always dependent on the RB resistor (the slower of the two discharge/charge cycles.

3. Pulse Width occurs when Q2 is off.

   (a) Analyzing the circuit, observe that when C1 charges to 0.7V through R3, Q2 will turn on. This means that C1 and R3 (the time it takes C1 to discharge from -4.3V and charge to 0.7V) will determine the Pulse Width or time off of Q2.

4. Pulse Space occurs when Q2 is on.

   (a) Analyzing the circuit, observe that when C2 charges to 0.7V through R1, Q1 will turn on which will force Q2 off. This means that C2 and R1 (the time it takes C2 to discharge from -4.3V and charge to 0.7V) will determine the Pulse Space or time off of Q2.

5. Capacitor Formulas Derived:

   (a) $VC = V_{fin} - (V_{fin} - V_{in})e^{\frac{-t}{RC}}$

   When V_{C1} reaches 0.7V, the output transitions from a high to a low.

   - $0.7V = V_{Fin} - (V_{fin} - V_{in})e^{\frac{-t}{RC}}$

     If the transistor could charge uninterrupted, it would charge to V_{CC} in this example 5V.
   - $0.7V = 5V - (5V - V_{in})e^{\frac{-t}{RC}}$

The tricky part is that we must account for the changing polarity of the capacitor voltage, meaning the capacitor has to discharge down from -4.3V to 0 and up to 0.7V.

- $0.7V = 5V - (5V - (-4.3V))e^{\frac{-t}{RC}}$
  - $0.7V - 5V = -(5V - (-4.3V))e^{\frac{-t}{RC}}$
  - $-4.3V = -(5V + 4.3V)e^{\frac{-t}{RC}}$
  - $-4.3V = -(9.3V)e^{\frac{-t}{RC}}$
  - $\frac{-4.3V}{-9.3V} = e^{\frac{-t}{RC}}$
  - $\frac{4.3}{9.3} = e^{\frac{-t}{RC}}$
  - $LN\frac{4.3}{9.3} = \frac{-t}{RC}$
  - $-0.7714 = \frac{-t}{RC}$
  - $C = \frac{t}{0.7714R}$

    In our example, $R = RB_{Q2} = R3$ and $C = C1$.
  - $C1 = \frac{t}{0.7714(43K\Omega)}$

Select a time or desired frequency. For this example, we will use 1Khz.

- $t = \frac{1}{2f}$
  - $t = \frac{1}{2(1Khz)}$
  - $t = 500\mu S$

Solve for C1.

- $C1 = \frac{t}{0.7714(43K\Omega)}$
  - $C1 = \frac{500\mu S}{0.7714(43K\Omega)}$
  - ✓ $C1 = 15.074nF$

Assuming a desired 50% duty cycle if the two RBs (R1 and R3) are equal, C2 will be equal to C1.

- $C2 = C1 = 15.074nF$

11c.4.3 Capacitor Value Approximation Method

The Approximation Method assumes that the discharge voltage is down to 0V and the charge voltage is VCC.

1. $VC = V_{fin} - (V_{fin} - V_{in})e^{\frac{-t}{RC}}$

   - $0V = VCC - (VCC - (-VCC))e^{\frac{-t}{RC}}$
   - $-VCC = -(VCC - (-VCC))e^{\frac{-t}{RC}}$
   - $-VCC = -(VCC + VCC)e^{\frac{-t}{RC}}$
   - $-VCC = -(2VCC)e^{\frac{-t}{RC}}$
   - $\frac{-VCC}{-2VCC} = e^{\frac{-t}{RC}}$

- $\frac{1}{2} = e^{\frac{-t}{RC}}$
- $LN\frac{1}{2} = \frac{-t}{RC}$
- $-0.693147 = \frac{-t}{RC}$
- $C = \frac{-t}{-0.693147R}$
- $C = \frac{t}{0.693147R}$

✓ $C = \frac{PW}{0.693147R}$ (Substitute desired Pulse Width for time)

Design for previous circuit at 1Khz:

- ○ $C = \frac{500\mu S}{0.693147(43K\Omega)}$
- ○ $C = \frac{500\mu S}{0.693147(43K\Omega)}$
- ○ $C = 16.776nF$

$C1$ & $C2 = 16.776nF$ (using the approximation method)

11c.4.4 Improved Rise Time Astable Variant

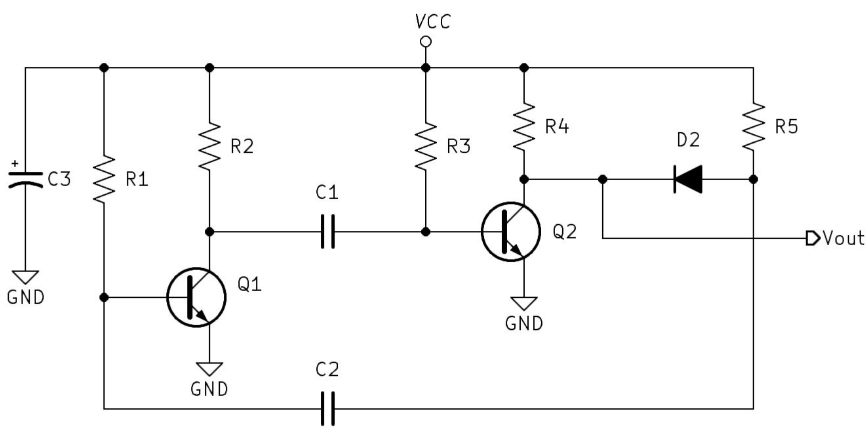


Figure 11c.3: Improved Rise Time BJT Astable Variant

R5 Calculations for Improved Rise Time BJT Astable Variant

1. Pulse Width $(PW) = 5\tau$
   (a) $\tau = \frac{PW}{5}$
2. $\tau = RC$
3. Substitute the formulas in terms of τ.
   (a) $RC = \frac{PW}{5}$
   (b) $R5 \times C2 = \frac{PW}{5}$
   ✓ $R5 = \frac{PW}{C2 \times 5}$

11c.5 BJT Monostable Multivibrator:

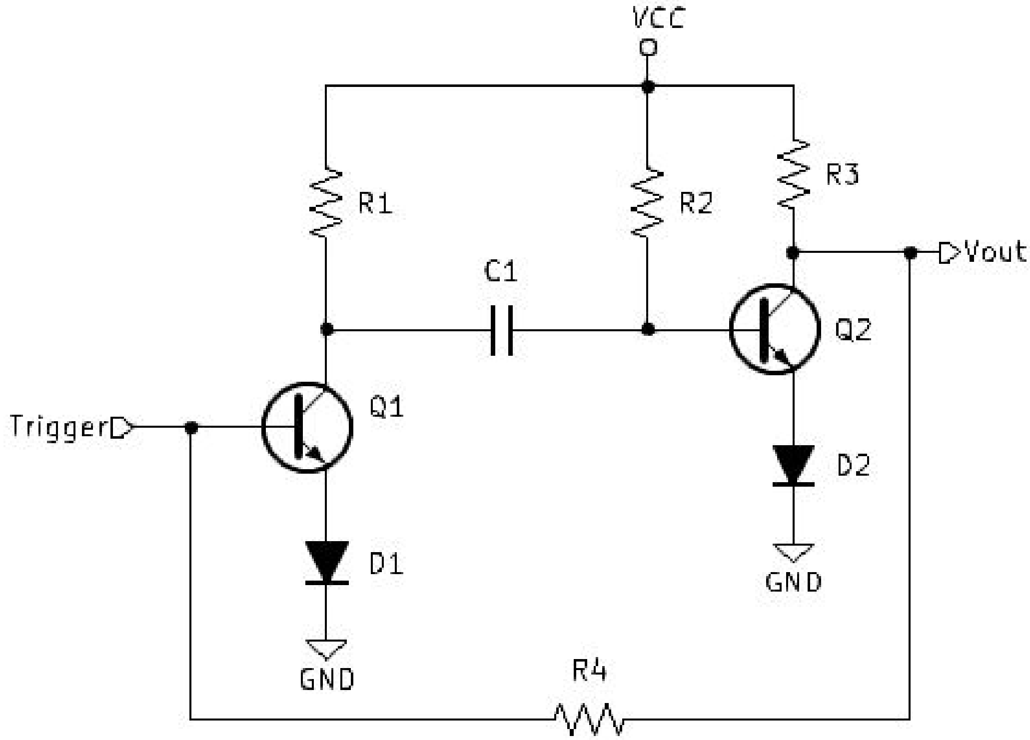


Figure 11c.4: Bipolar Junction Transistor Monostable Multivibrator

11c.5.1 Introduction

A Bipolar Junction Transistor (BJT) monostable multivibrator, also known as a one-shot multivibrator, is a pivotal circuit in electronics designed to generate a single output pulse of a specific duration in response to an external trigger. Unlike its astable counterpart, the BJT monostable multivibrator has one stable state and one unstable state. Upon receiving a triggering signal, the circuit temporarily shifts to its unstable state, producing a pulse before reverting to its stable state. This functionality is achieved using a combination of BJTs, resistors, and capacitors, which determine the pulse width. Monostable multivibrators are extensively used in applications such as pulse generation, timers, and debouncing switches, offering precise control over timing events in various electronic systems.

11c.5.2 Calculating Resistor Values

1. The transistors will be like switches, designed to operate in the Saturation and Cutoff regions.
2. Choose a saturation current IC_{Sat}.
3. $R1$ & $R3 = \frac{VCC - VF_{Diode}}{IC_{Sat}}$
4. $IB_{Sat} = \frac{IC_{Sat}}{Beta_{min}}$
5. $R2$ & $R4 = \frac{VCC-(V_{BE}+VF_{Diode})}{IB_{Sat}}$
6. Example:
   - Given: $VCC = 5V$, $IC_{Sat} = 10mA$, and $Beta_{min} = 100$
     - $R1$ & $R3 = 430\Omega$
     - $R2$ & $R4 = 36K\Omega$

11c.5.3 Calculating the Capacitor Value

- Diodes D1 and D2 serve two purposes. They raise the trigger threshold voltage which helps prevent false triggers and protect the transistors from negative over-voltage/current at the base to emitter junction.

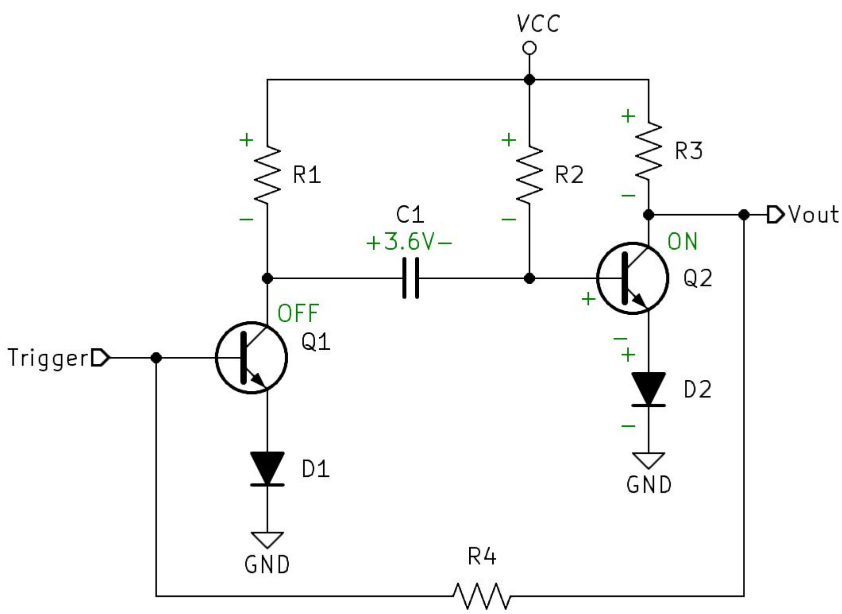


Figure 11c.5: Bipolar Junction Transistor Monostable Multivibrator Stable State

- **Stable State** Observe Figure11c.5. Q2 will immediately turn on due to the unrestricted base resistor R2. The current path for Q1 to turn on goes through R3 and R4. This means Q2 will turn on faster than Q1 and once Q1 is on it will take away VCC leaving only the 0.7 volts of D2 which will not be enough to bias on both Q1 and D1. This is the happy or stable state of the monostable and Vout is low.

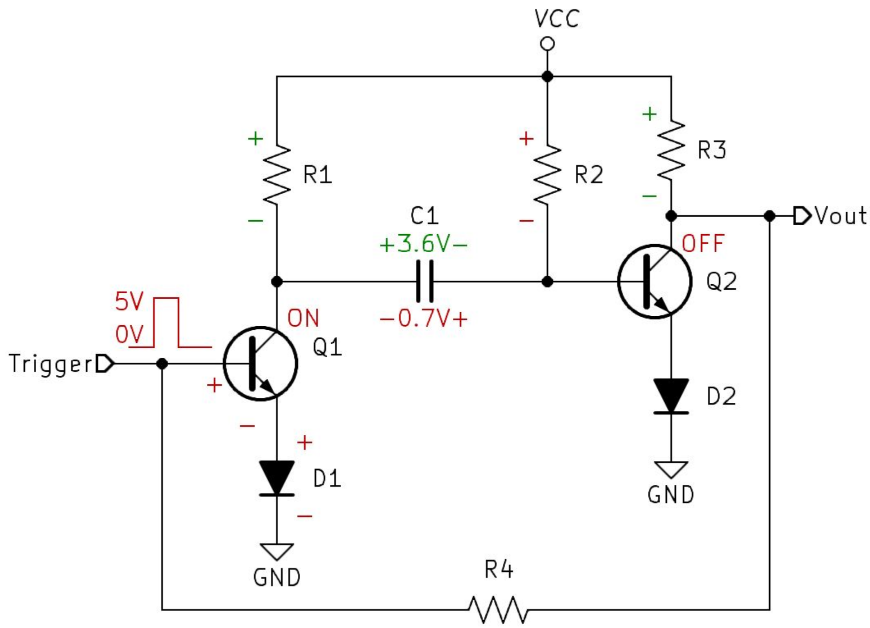


Figure 11c.6: Bipolar Junction Transistor Monostable Multivibrator Un-Stable State

- To drive the monostable output high, a positive going trigger signal must be applied to the base of Q1. Ideally, the trigger pulse width is significantly less than the desired pulse width of the Monostable Multivibrator.

- Observe Figure 11c.6. As Q1 is biased on from the Trigger signal, Q2 initially will see -3.6V at its base which will turn Q2 off. C1 will begin to discharge the -3.6V through the path of R2, Q1, and D1. Once discharged Q2 will begin to charge and once it reaches 0.7V, the base of Q2 will now have 1.4V (VC1 +VD1) this will turn on Q2 which will then force Q1 back off returning the circuit to return to its stable state.

- $VC = V_{fin} - (V_{fin} - V_{in})e^{\frac{-t}{RC}}$
  - $0.7V = VCC - (VCC - (-(VCC - 1.4V))e^{\frac{-PW}{R2C1}}$
  - $0.7V = VCC - (VCC + VCC - 1.4V)e^{\frac{-PW}{R2C1}}$
  - $0.7V = VCC - (2VCC - 1.4V)e^{\frac{-PW}{R2C1}}$
  - $0.7V - VCC = -(2VCC - 1.4V)e^{\frac{-PW}{R2C1}}$
  - $0.7V - VCC = (-2VCC + 1.4V)e^{\frac{-PW}{R2C1}}$
  - $\frac{0.7V - VCC}{-2VCC + 1.4V} = e^{\frac{-PW}{R2C1}}$

- ○ $\frac{VCC-0.7V}{2VCC-1.4V} = e^{\frac{-PW}{R2C1}}$
- ○ $\frac{1(VCC-0.7V)}{2(VCC-0.7V)} = e^{\frac{-PW}{R2C1}}$
- ○ $\frac{1}{2} = e^{\frac{-PW}{R2C1}}$
- ○ $LN\frac{1}{2} = \frac{-PW}{R2C1}$
- ○ $-0.693 = \frac{-PW}{R2C1}$
- ○ $C1 = \frac{-PW}{R2\times(-0.693)}$
- ✓ $C1 = \frac{PW}{R2\times 0.693}$

Example:

Given: $VCC = 5V$, $IC_{Sat} = 10mA$, $Beta_{min} = 100$, R1 & $R3 = 500\Omega$, $R2$ & $R4 = 43K\Omega$, and $PW = 1\ Second$.

Find: $C1$

- ○ $C1 = \frac{PW}{R2\times 0.693}$
- ○ $C1 = \frac{-1Sec}{43K\Omega\times 0.693}$
- ✓ $C1 = 33.558\mu F$

11c.5.4 Trigger Circuit

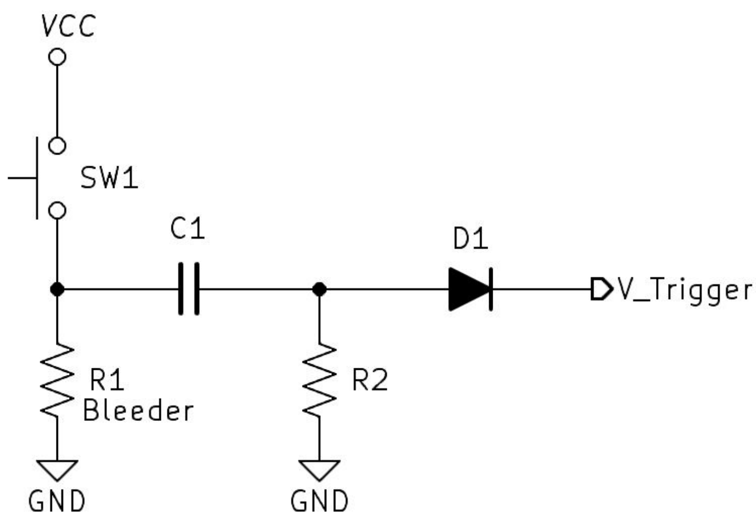


Figure 11c.7: Trigger Circuit

The trigger signal's pulse width needs to be significantly less than the pulse width of the monostable multivibrator ideally, $PW_{Trigger} \leq \frac{PW_{MonoStable}}{10}$. A differentiated pulse signal in Figure 11c.7. will eliminate switch bounce and will prevent unintended re-triggers of the Monostable Multivibrator.

Trigger Circuit Setup:

- Consider the desired Monostable Multivibrator output Pulse Width.
- Design an RC Differentiated circuit (C1 and R2) to produce a pulse that is ten times smaller than the desired Monostable Pulse Width, $PW_{Trigger} \leq \frac{PW_{MonoStable}}{10}$. Review RC Circuits Differentiators on page 65 if necessary.
- Make R1 the bleeder resistor ten times larger than R2, $R_{Bleeder} \approx 10 \times R2$.
- Diode D1 clips the negative spike of the differentiated signal.
- Build and test. If the signal looks correct but the Monostable Multivibrator will not trigger, raise the resistance of R2 until the Monostable is reliably triggered by switch SW1.

11c.6 Extreme Duty Cycle Astable Multivibrator Circuit Variant

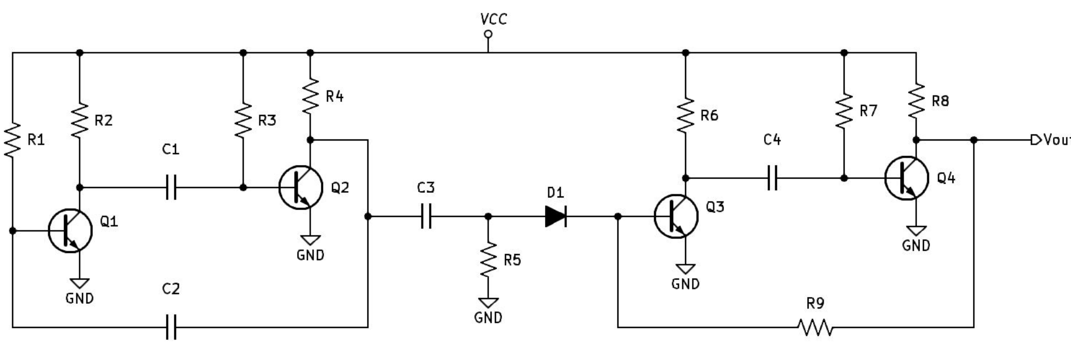


Figure 11c.8: Astable Multivibrator Extreme Duty Cycle Circuit Variant

11c.6.1 Introduction

An extreme duty cycle astable multivibrator circuit variant is a specialized form of the conventional astable multivibrator designed to generate output pulses with exceptionally high or low duty cycles. Unlike standard configurations, which typically aim for a 50% duty cycle or moderately skewed pulse widths, this variant uses tailored resistor and capacitor values to produce output signals where the on-time or off-time significantly dominates the cycle. Such circuits are particularly useful in applications requiring precise control over pulse duration relative to the overall period, such as in pulse-width modulation (PWM) for power control, specialized timing circuits, and signal modulation tasks. By leveraging components like Bipolar Junction Transistors (BJTs) or integrated circuits (ICs), these multivibrators can achieve the desired extreme duty cycles while maintaining stability and reliability in various electronic systems.

- The standard Astable Multivibrator circuit likes to operate around 50% duty cycle.
- To produce an extreme duty cycle waveform,an a standard Astable circuit can be used to drive/trigger a standard Monostable circuit as seen in Figure 11c.8.
- Design steps:
  1. Design a 50% duty cycle BJT Astable to operate at the desired frequency (Q1 & Q2).
  2. Design a BJT Monostable circuit to produce the desired pulse width (Q3 & Q4).
  3. Use an RC circuit to differentiate the output of the Astable to produce the trigger pulse for the Monostable circuit (C3 & R5) ensuring that the differentiated waveform's pulse width is significantly less than the desired output pulse width.
  4. The Diode D1 protects the base of Q3 from any negative voltage and keeps R6 isolated from the monostable biasing circuit.

11c.7 555 Timer Astable Multivibrator:

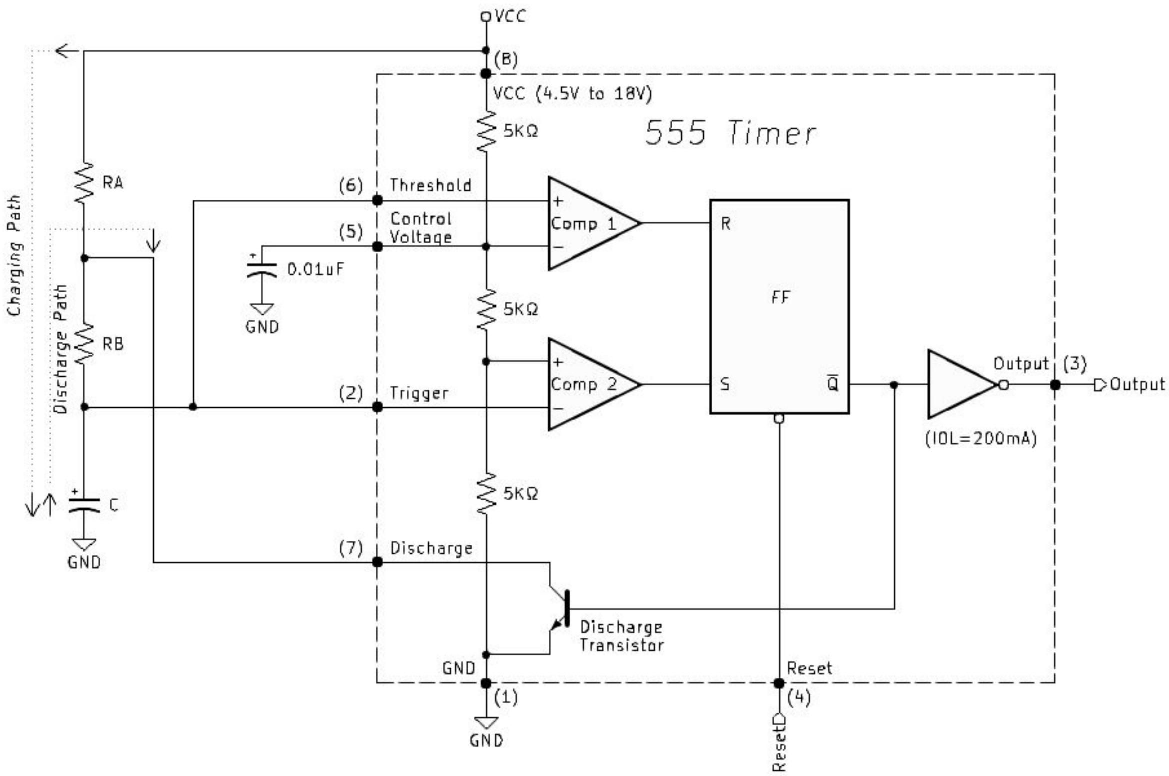


Figure 11c.9: Simplified block diagram of a 555 timer with the external timer components to form an astable multivibrator - Kleitz [23]

11c.7.1 Introduction

The 555 timer astable multivibrator is a versatile and widely-used electronic circuit configuration that generates a continuous square wave output. At the heart of this configuration is the 555 timer IC, a highly reliable and easy-to-use integrated circuit that has become a staple in electronics design. In its astable mode, the 555 timer operates without any stable states, constantly oscillating between high and low output levels to produce a periodic waveform. This oscillation is achieved by carefully selecting external resistors and capacitors, which determine the frequency and duty cycle of the output signal. The 555 timer astable multivibrator is commonly employed in applications such as clock pulse generation, LED flashers, and tone generators, offering a simple yet powerful solution for creating precise and adjustable timing signals in a variety of electronic projects.

11c.7.2 Circuit Analysis

- Internally, VCC is divided across three 5KΩ resistors. Comparator 1 will have $\frac{2VCC}{3}$ on its minus input. Comparator 2 will have $\frac{1VCC}{3}$ on its plus input.
- When power is initially turned on, the capacitor voltage VC, starting at zero volts will charge up through RA and RB. Starting off, Comparator 1 will rail negative, and Comparator 2 will rail positive. This will set the Flip Flop, $\overline{Q}$ will equal a Low, the discharge transistor is off and the output, after the inverter, is High.
- As the capacitor voltage charges up through RA and RB, it will eventually cross the $\frac{1VCC}{3}$ threshold, this will cause Comparator 2 to rail negative which will cause the Flip Flow to enter a (0,0) Hold state (no change at the output).
- The capacitor voltage continues to charge through RA and RB and will eventually cross the $\frac{2VCC}{3}$ threshold. At this point, Comparator 1 will rail positive resulting in a reset for the Flip Flop, $\overline{Q}$ will go high and turn on the discharge transistor. Additionally, the output will go low via the inverter.
- Once the discharge transistor is on, ground potential is essentially placed between RA and RB. This will cause the capacitor to start to discharge through RB.
- As VC discharges and falls below $\frac{2VCC}{3}$, Comparator 1 will rail low which will cause the Flip Flop to go into a hold state (0,0).
- As VC continues to discharge and falls below $\frac{1VCC}{3}$, Comparator 2 will rail high and cause the Flip Flop to go into a Set state. $\overline{Q}$ will go low which will turn off the discharge transistor and the output will go high via the inverter.
- From the analysis, we can see that our Astable output will oscillate as VC discharges and charges between $\frac{2VCC}{3}$ and $\frac{1VCC}{3}$.

11c.7.3 Pulse Width (Charge Time) Formula Derivation

During the charge cycle, we know that the output is high and the capacitor will charge through RA and RB from $\frac{1VCC}{3}$ to $\frac{2VCC}{3}$.

- $VC = V_{fin} - (V_{fin} - V_{in})e^{\frac{-t}{RC}}$
  - $\frac{2VCC}{3} = VCC - (VCC - \frac{1VCC}{3})e^{\frac{-PW}{(RA+RB)C}}$
  - $\frac{2VCC}{3} = \frac{3VCC}{3} - (\frac{3VCC}{3} - \frac{1VCC}{3})e^{\frac{-PW}{(RA+RB)C}}$
  - $\frac{2VCC}{3} = \frac{3VCC}{3} - (\frac{2VCC}{3})e^{\frac{-PW}{(RA+RB)C}}$
  - $\frac{2VCC}{3} - \frac{3VCC}{3} = -(\frac{2VCC}{3})e^{\frac{-PW}{(RA+RB)C}}$
  - $\frac{-1VCC}{3} = (\frac{-2VCC}{3})e^{\frac{-PW}{(RA+RB)C}}$

- $\frac{-1VCC}{3} \times \frac{-3}{2VCC} = e^{\frac{-PW}{(RA+RB)C}}$
- $\frac{3VCC}{6VCC} = e^{\frac{-PW}{(RA+RB)C}}$
- $\frac{1}{2} = e^{\frac{-PW}{(RA+RB)C}}$
- $LN\frac{1}{2} = \frac{-PW}{(RA+RB)C}$
- $-0.693147 = \frac{-PW}{(RA+RB)C}$
- ✓ $PW = 0.693147(RA + RB)C$

11c.7.4 Pulse Space (Discharge Time) Formula Derivation

During the discharge cycle, we know that the output is low and the capacitor will discharge through RB from $\frac{2VCC}{3}$ to $\frac{1VCC}{3}$.

- $VC = V_{fin} - (V_{fin} - V_{in})e^{\frac{-t}{RC}}$
  - $\frac{1VCC}{3} = 0 - (0 - \frac{2VCC}{3})e^{\frac{-PS}{(RB)C}}$
  - $\frac{1VCC}{3} = -(-\frac{2VCC}{3})e^{\frac{-PS}{(RB)C}}$
  - $\frac{1VCC}{3} = (\frac{2VCC}{3})e^{\frac{-PS}{(RB)C}}$
  - $\frac{1VCC}{3} = (\frac{2VCC}{3})e^{\frac{-PS}{(RB)C}}$
  - $\frac{1VCC}{3} \times \frac{3}{2VCC} = e^{\frac{-PS}{(RB)C}}$
  - $\frac{3VCC}{6VCC} = e^{\frac{-PS}{(RB)C}}$
  - $\frac{1}{2} = e^{\frac{-PS}{(RB)C}}$
  - $LN\frac{1}{2} = \frac{-PS}{(RB)C}$
  - $-0.693147 = \frac{-PS}{(RB)C}$
  - ✓ $PS = 0.693147(RB)C$

11c.7.5 Design Example

- Pick a desired frequency. For this example, we will use 1Khz.
  - $Period = \frac{1}{frequency} = \frac{1}{1Khz} = 1mS$
  - $Period = PW + PS$
    - $Period = 0.693147(RA + RB)C + 0.693147(RB)C$
    - $Period = 0.693147C[(RA + RB) + (RB)]$
    - $Period = 0.693147C[RA + 2RB]$
    - $\frac{Period}{0.693147C} = [RA + 2RB]$

* $\frac{1mS}{0.693147C} = [RA + 2RB]$
  Pick a capacitor value ($1\mu F$)
* $\frac{1mS}{0.693147 \times 1\mu F} = RA + 2RB$
* $1.443 \times 10^3 = RA + 2RB$
  Make RA approximately $\frac{1}{3}$ of the total resistance.
* $RA \approx \frac{1.443 \times 10^3}{3} \approx 480.898\Omega$
  Round RA to a standard value

✓ $RA = 470\Omega$
  Solve for RB
* $2RB = 1.443K\Omega - 470\Omega$
* $2RB = 972.695\Omega$
* $RB = \frac{972.695\Omega}{2}$

✓ $RB = 486.348\Omega$ ($470\Omega + 18\Omega$)

11c.7.6 Adjustable Duty Cycle Circuit

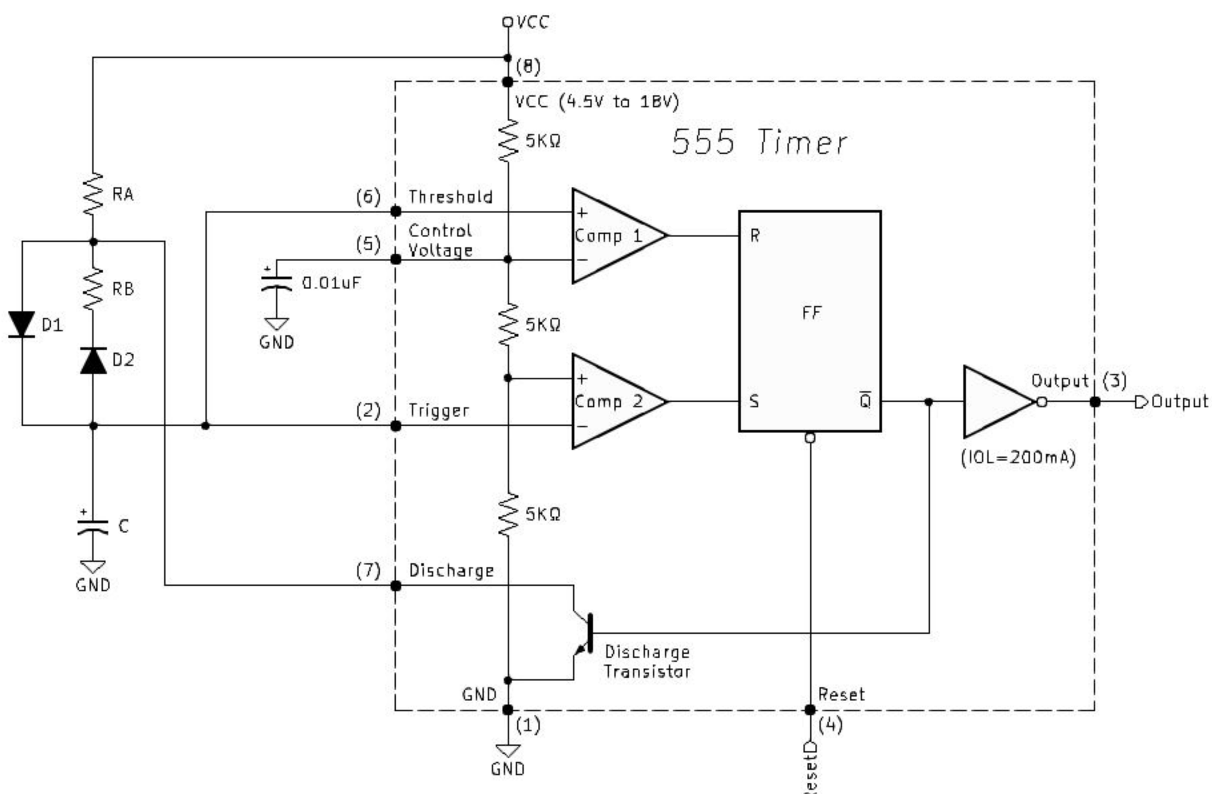


Figure 11c.10: Adjustable Duty Cycle 555 Timer Astable Circuit

The Adjustable Duty Cycle Circuit of Figure 11c.10 will operate similarly to the standard 555 Timer Astable Multivibrator circuit previously analyzed. The main difference is that the charge path, which determines the Pulse Width of the output, is now through RA and D1 and the discharge path is through D2 and RB. Technically, D2 is not necessary however, it is useful when designing for 50% duty cycle outputs to balance the charge and discharge paths allowing for the RA and RB resistances to be equal value.

11c.7.7 Pulse Width (Charge Time) Formula Derivation

During the charge cycle, we know that the output is high and the capacitor will charge through RA and D1 from $\frac{1VCC}{3}$ to $\frac{2VCC}{3}$.

- $VC = V_{fin} - (V_{fin} - V_{in})e^{\frac{-t}{RC}}$
  - $\frac{2VCC}{3} = (VCC - VD1) - ((VCC - VD1) - \frac{1VCC}{3})e^{\frac{-PW}{RA \times C}}$
  - $\frac{2VCC}{3} - (VCC - VD1) = -((VCC - VD1) - \frac{1VCC}{3})e^{\frac{-PW}{RA \times C}}$
  - $\frac{2VCC}{3} - (VCC - VD1) = (-(VCC - VD1) + \frac{1VCC}{3})e^{\frac{-PW}{RA \times C}}$
  - $\frac{2VCC}{3} - (VCC - VD1) = (-VCC + VD1 + \frac{1VCC}{3})e^{\frac{-PW}{RA \times C}}$
  - $\frac{2VCC}{3} - (\frac{3VCC}{3} - \frac{3VD1}{3}) = (-\frac{3VCC}{3} + \frac{3VD1}{3} + \frac{1VCC}{3})e^{\frac{-PW}{RA \times C}}$
  - $\frac{2VCC}{3} - \frac{3VCC}{3} + \frac{3VD1}{3} = (-\frac{3VCC}{3} + \frac{3VD1}{3} + \frac{1VCC}{3})e^{\frac{-PW}{RA \times C}}$
  - $\frac{2VCC - 3VCC + 3VD1}{3} = (\frac{-3VCC + 3VD1 + 1VCC}{3})e^{\frac{-PW}{RA \times C}}$
  - $\frac{-1VCC + 3VD1}{3} = (\frac{-2VCC + 3VD1}{3})e^{\frac{-PW}{RA \times C}}$
  - $\frac{-1VCC + 3VD1}{3} \times \frac{3}{-2VCC + 3VD1} = e^{\frac{-PW}{RA \times C}}$
  - $\frac{-1VCC + 3VD1}{-2VCC + 3VD1} = e^{\frac{-PW}{RA \times C}}$
  - $\frac{-1VCC + 3(0.7)}{-2VCC + 3(0.7)} = e^{\frac{-PW}{RA \times C}}$
  - $\frac{-1VCC + 2.1}{-2VCC + 2.1} = e^{\frac{-PW}{RA \times C}}$
  - $\frac{1VCC - 2.1}{2VCC - 2.1} = e^{\frac{-PW}{RA \times C}}$
  - $LN[\frac{1VCC - 2.1}{2VCC - 2.1}] = \frac{-PW}{RA \times C}$
  - $LN[\frac{VCC - 2.1}{2VCC - 2.1}] \times (RA \times C) = -PW$
  - ✓ $PW = -[(LN[\frac{VCC - 2.1}{2VCC - 2.1}]) \times (RA \times C)]$

11c.7.8 Pulse Space (Discharge Time) Formula Derivation

During the discharge cycle, we know that the output is low and the capacitor will discharge through RB and D2 from $\frac{2VCC}{3}$ to $\frac{1VCC}{3}$.

- $VC = V_{fin} - (V_{fin} - V_{in})e^{\frac{-t}{RC}}$
  - $\frac{1VCC}{3} = VD2 - (VD2 - \frac{2VCC}{3})e^{\frac{-PS}{(RB)C}}$
  - $\frac{1VCC}{3} - VD2 = -(VD2 - \frac{2VCC}{3})e^{\frac{-PS}{(RB)C}}$
  - $\frac{1VCC}{3} - \frac{3VD2}{3} = (-\frac{3VD2}{3} + \frac{2VCC}{3})e^{\frac{-PS}{(RB)C}}$
  - $\frac{1VCC-3VD2}{3} = (\frac{-3VD2+2VCC}{3})e^{\frac{-PS}{(RB)C}}$
  - $\frac{1VCC-3VD2}{3} \times \frac{3}{-3VD2+2VCC} = e^{\frac{-PS}{(RB)C}}$
  - $\frac{1VCC-3VD2}{-3VD2+2VCC} = e^{\frac{-PS}{(RB)C}}$
  - $\frac{1VCC-3VD2}{2VCC-3VD2} = e^{\frac{-PS}{(RB)C}}$
  - $\frac{1VCC-3(0.7)}{2VCC-3(0.7)} = e^{\frac{-PS}{(RB)C}}$
  - $\frac{1VCC-2.1}{2VCC-2.1} = e^{\frac{-PS}{(RB)C}}$
  - $LN[\frac{1VCC-2.1}{2VCC-2.1}] = \frac{-PS}{(RB)C}$
  - $LN[\frac{VCC-2.1}{2VCC-2.1}] \times (RB \times C) = -PS$
  - ✓ $PS = -[(LN[\frac{VCC-2.1}{2VCC-2.1}]) \times (RB \times C)]$
    * Notice the Pulse Space formula is the same formula as the Pulse Width formula the only difference is the substitution of RB for RA.

11c.7.9 Design Example 1

Given:

Design an Adjustable Duty Cycle 555 Timer Astable Circuit to produce a 10Khz 50% Duty Cycle signal using 5 volts for VCC and a 0.1μF capacitor.

Find:

- RA
- RB

Solve:

- **Solve for RA:**
  - $PW = -[(LN[\frac{VCC-2.1}{2VCC-2.1}]) \times (RA \times C)]$
  - $\frac{PW}{(LN[\frac{VCC-2.1}{2VCC-2.1}])\times C} = -(RA)$
  - $RA = \frac{-PW}{(LN[\frac{VCC-2.1}{2VCC-2.1}])\times C}$
    - * $PW = \frac{1}{2F}$
      - $PW = \frac{1}{2\times 10Khz}$
      - $PW = 50\mu S$
  - $RA = \frac{-50\mu S}{(LN[\frac{5-2.1}{2(5)-2.1}])\times 0.1\mu F}$
  - $RA = \frac{-50\mu S}{(LN[\frac{2.9}{10-2.1}])\times 0.1\mu F}$
  - $RA = \frac{-50\mu S}{(LN[\frac{2.9}{7.9}])\times 0.1\mu F}$
  - $RA = \frac{-50\mu S}{(-1.00215)\times 0.1\mu F}$
  - $RA = \frac{-50\mu S}{-100.215\times 10^{-9}S}$
  - ✓ $RA = 498.927\Omega$

- **Solve for RB:**
  - $PS = -[(LN[\frac{VCC-2.1}{2VCC-2.1}]) \times (RB \times C)]$
  - $\frac{PS}{(LN[\frac{VCC-2.1}{2VCC-2.1}])\times C} = -(RB)$
  - $RB = \frac{-PS}{(LN[\frac{VCC-2.1}{2VCC-2.1}])\times C}$
    - * $PS = \frac{1}{2F}$
      - $PS = \frac{1}{2\times 10Khz}$
      - $PS = 50\mu S$

- $RB = \frac{-50\mu S}{(LN[\frac{5-2.1}{2(5)-2.1}])\times 0.1\mu F}$
- $RB = \frac{-50\mu S}{(LN[\frac{2.9}{10-2.1}])\times 0.1\mu F}$
- $RB = \frac{-50\mu S}{(LN[\frac{2.9}{7.9}])\times 0.1\mu F}$
- $RB = \frac{-50\mu S}{(-1.00215)\times 0.1\mu F}$
- $RB = \frac{-50\mu S}{-100.215\times 10^{-9}S}$

✓ $RB = 498.927\Omega$

11c.7.10 Design Example 2

Given:

Design an Adjustable Duty Cycle 555 Timer Astable Circuit to produce an output waveform with a Pulse Space equal to 1.5mS and a Pulse Width of 20mS using 5 volts for VCC and a 10μF capacitor.

Find:

- RA
- RB

Solve:

- **Solve for RA:**
  - $RA = \frac{-PW}{(LN[\frac{VCC-2.1}{2VCC-2.1}])\times C}$
  - $RA = \frac{-1.5mS}{(LN[\frac{5-2.1}{2(5)-2.1}])\times 10\mu F}$
  - $RA = \frac{-1.5mS}{(LN[\frac{2.9}{10-2.1}])\times 10\mu F}$
  - $RA = \frac{-1.5mS}{(LN[\frac{2.9}{7.9}])\times 10\mu F}$
  - $RA = \frac{-1.5mS}{(-1.00215)\times 10\mu F}$
  - $RA = \frac{-1.5mS}{-10.0215\times 10^{-6}S}$

  ✓ $RA = 149.678\Omega$

- **Solve for RB:**
  - $RB = \frac{-PS}{(LN[\frac{VCC-2.1}{2VCC-2.1}])\times C}$
  - $RB = \frac{-20mS}{(LN[\frac{5-2.1}{2(5)-2.1}])\times 10\mu F}$

- $RB = \frac{-20mS}{(LN[\frac{2.9}{10-2.1}])\times 10\mu F}$
- $RB = \frac{-20mS}{(LN[\frac{2.9}{7.9}])\times 10\mu F}$
- $RB = \frac{-20mS}{(-1.00215)\times 10\mu F}$
- $RB = \frac{-20mS}{-10.0215\times 10^{-6}S}$
- ✓ $RA = 1.996K\Omega$

11c.8 555 Timer Monostable Multivibrator:

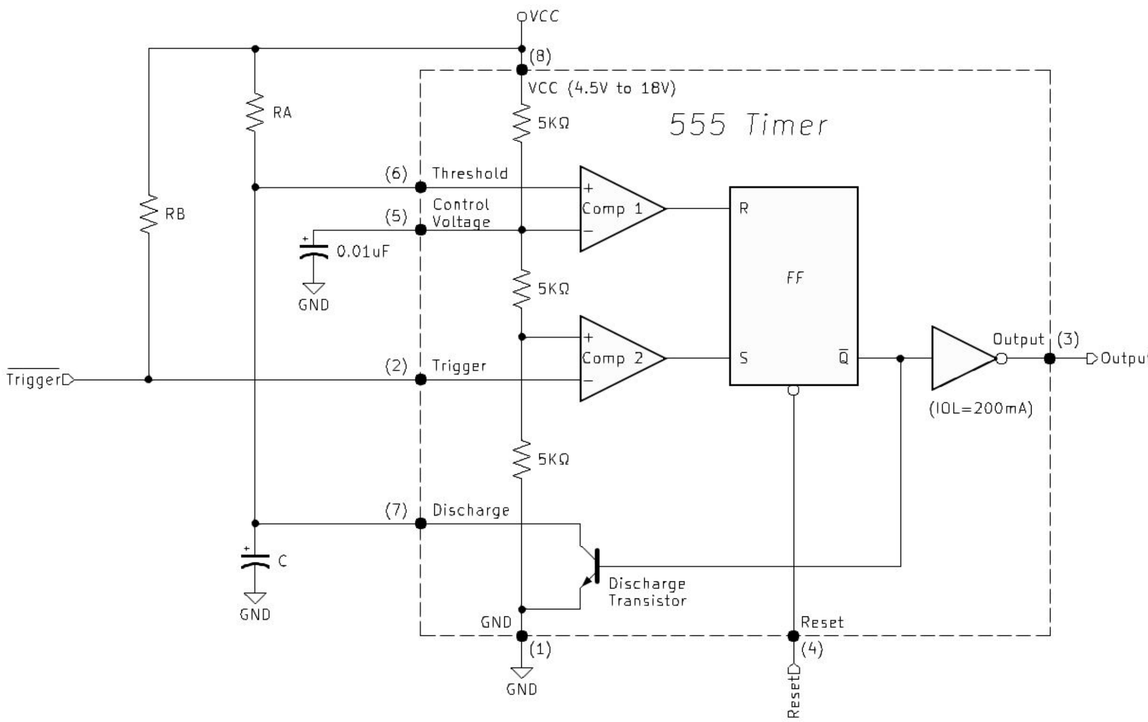


Figure 11c.11: Monostable Multivibrator circuit using a 555 Timer

11c.8.1 Introduction

The 555 timer monostable multivibrator is an essential electronic circuit configuration that generates a single, precise output pulse in response to an external trigger signal. Utilizing the versatile 555 timer IC, this monostable mode operation involves the circuit transitioning from its stable state to an unstable state upon receiving a trigger. The duration of the output pulse is determined by external components, typically a resistor and capacitor, allowing for adjustable and accurate timing. After the pulse is completed, the circuit automatically returns to its stable state, ready to be triggered again. The 555 timer monostable multivibrator is widely used in applications such as pulse generation, timing delays, and debouncing switches, providing a reliable and straightforward solution for creating controlled timing events in various electronic systems.

11c.8.2 Circuit Analysis

- Internally, VCC is divided across three 5KΩ resistors. Comparator 1 will have $\frac{2VCC}{3}$ on its minus input. Comparator 2 will have $\frac{1VCC}{3}$ on its plus input.
- Assuming that the 555 timers internal Flip Flop power up in the Set condition, the output will be high, the Discharge Transistor will be off, and the Capacitor voltage will begin to charge through RA. Initially, Comparitor 1 will rail negative and Comparitor 2 will rail negative resulting in a Hold state for the Flip Flop.
- As the capacitor voltage charges, or increases, through RA it will eventually cross the $\frac{2VCC}{3}$ threshold. This will cause Comparitor 1 to rail positive and will result in a Reset condition for the Flip Flop, $\overline{Q}$ will go high turning on the Discharge Transistor and the Output will go low via the inverter.
- With the Discharge Transistor on, ground is placed across the capacitor. This will cause the capacitor to discharge rapidly to 0V. Comparator 1 will rail low with Comparator 2 already low, the Flip Flop to enter a Hold state.
- (Stable State) The output is now low and the only thing that will change the output is a $\overline{Trigger}$ signal on pin 2.
- Note, the $\overline{Trigger}$ pulse needs to be shorter, or faster, than the desired pulse width of the Monstables output.

11c.8.3 Pulse Width Calculations

- $VC = V_{fin} - (V_{fin} - V_{in})e^{\frac{-t}{RC}}$
  - $VC = \frac{2VCC}{3}$
  - $V_{fin} = VCC$
  - $V_{in} = 0V$
  - $t = PW$
  - $R = RA$
  - $C = C$
- $\frac{2VCC}{3} = VCC - (VCC - 0)e^{\frac{-PW}{RA \times C}}$
  - $\frac{2VCC}{3} - VCC = -(VCC)e^{\frac{-PW}{RA \times C}}$
  - $\frac{2VCC}{3} - \frac{3VCC}{3} = (-VCC)e^{\frac{-PW}{RA \times C}}$

- $\frac{-1VCC}{3} = (-VCC)e^{\frac{-PW}{RA \times C}}$
- $\frac{-1VCC}{3(-VCC)} = e^{\frac{-PW}{RA \times C}}$
- $\frac{-1VCC}{-3VCC} = e^{\frac{-PW}{RA \times C}}$
- $\frac{1}{3} = e^{\frac{-PW}{RA \times C}}$
- $LN(\frac{1}{3}) = \frac{-PW}{RA \times C}$
- $-1.0986 = \frac{-PW}{RA \times C}$
- $-1.0986 \times (RA \times C) = -PW$
- ✓ $PW = 1.0986(RA \times C)$

11c.8.4 Design Example 1

Design a Monostable Multivibrator Circuit using a 555 Timer to produce a 5-second pulse when triggered.

Given:

- $VCC = 5V$
- $RB = 10K\Omega$
- $C = 100\mu F$
- $PW = 5\ seconds$

Find:

- RA

Solve:

- $PW = 1.0986(RA \times C)$
  - $\frac{PW}{1.0986 \times C} = (RA)$
  - $RA = \frac{PW}{1.0986 \times C}$
- $RA = \frac{PW}{1.0986 \times C}$
  - $RA = \frac{5sec}{1.0986 \times 100\mu F}$
  - $RA = \frac{5sec}{1.0986 \times 10^{-6}}$
  - ✓ $RA = 45.512K\Omega$

11c.8.5 Trigger Circuits

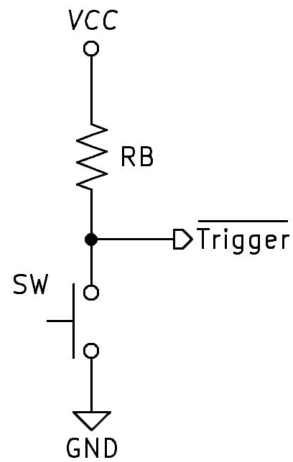


Figure 11c.12: Monostable Multivibrator Trigger Circuit 1, the monostable will see the switch bounce.

The Trigger Circuit Figure 11c.12. can be used to trigger a 555 Timer Monostable circuit with an output Pulse Width greater than or equal to ten times larger than the switch bounce.

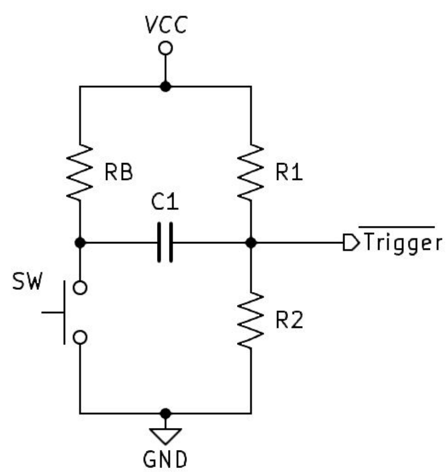
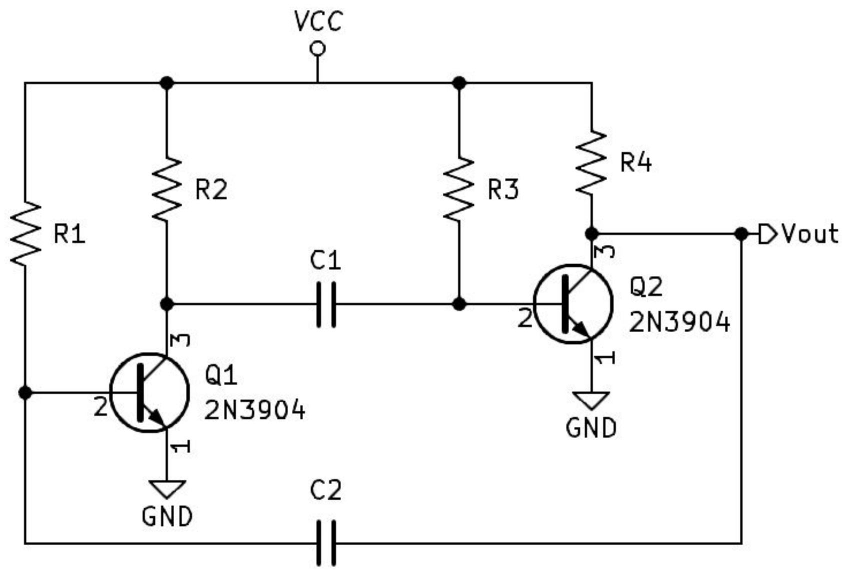


Figure 11c.13: Monostable Multivibrator Trigger Circuit 2, no switch bounce and trigger pulse-width can be adjusted

The Trigger Circuit Figure 11c.13. can be used to trigger a 555 Timer Monostable circuit with an output Pulse Width less than or equal to ten times larger than the switch bounce.

11c.9 Practice Questions:

1. Design a BJT Astable Multivibrator.



Given: $VCC = 18V$, $IC_{sat} = 30mA$, $PRF = 15Khz$, $DC = 58\%$, $Beta_{Q1} = 30$, $Beta_{Q2} = 50$.

Find: $R1$, $R2$, $R3$, $R4$, $C1$, $C2$.

Solve:

$R1 =$ ____________

$R2 =$ ____________

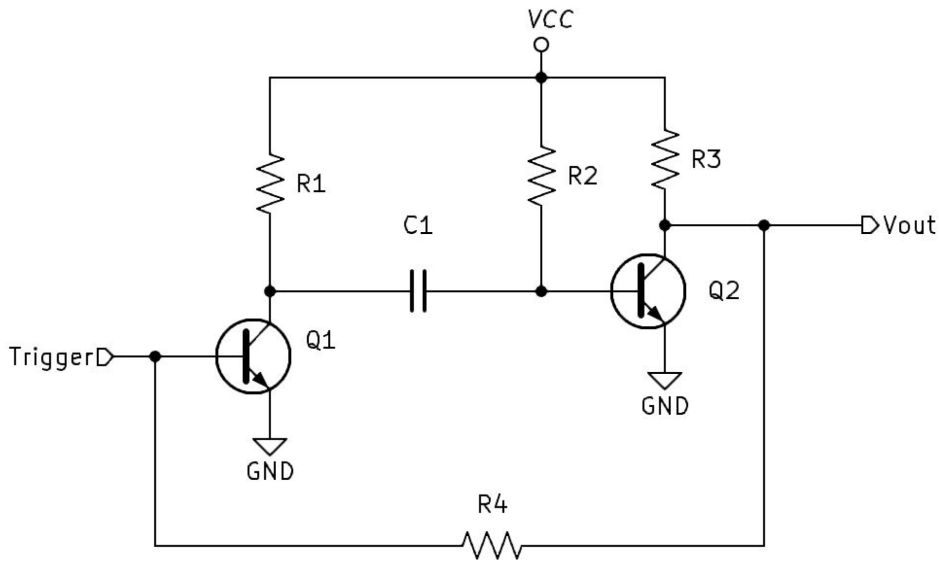
$R3 =$ ____________

$R4 =$ ____________

$C1 =$ ____________

$C2 =$ ____________

2. Design a BJT Monostable Multivibrator.



Given: $VCC = 10V$, $IC_{sat} = 10mA$, $PW = 2.00S$, $Beta_{Q1} = 120$, $Beta_{Q2} = 60$ (Make $IB = 2 \times IB_{min}$).

Find: $R1$, $R2$, $R3$, $R4$, $C1$, $C2$.

Solve:

$R1 =$ ____________

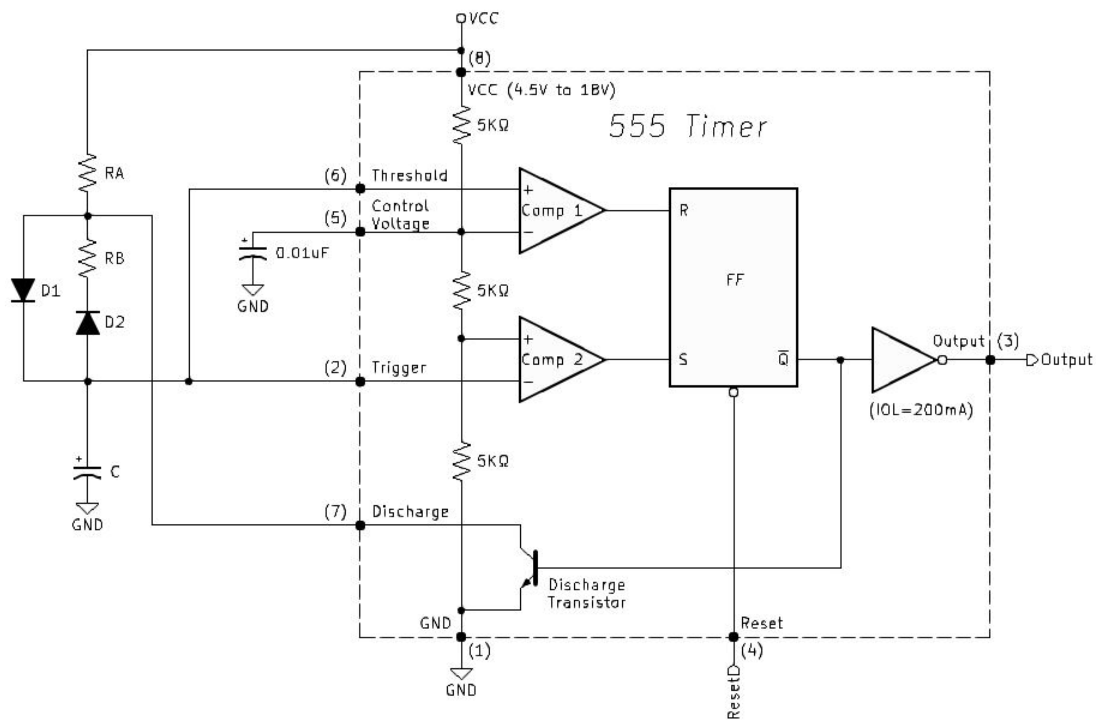
$R2 =$ ____________

$R3 =$ ____________

$R4 =$ ____________

$C1 =$ ____________

3. Design an Astable Multivibrator using a 555 Timer.



Given: $VCC = 10V$, $PW = 38.443\mu S$, $PS = 1.636\mu S$. $C1 = 0.001\mu F$

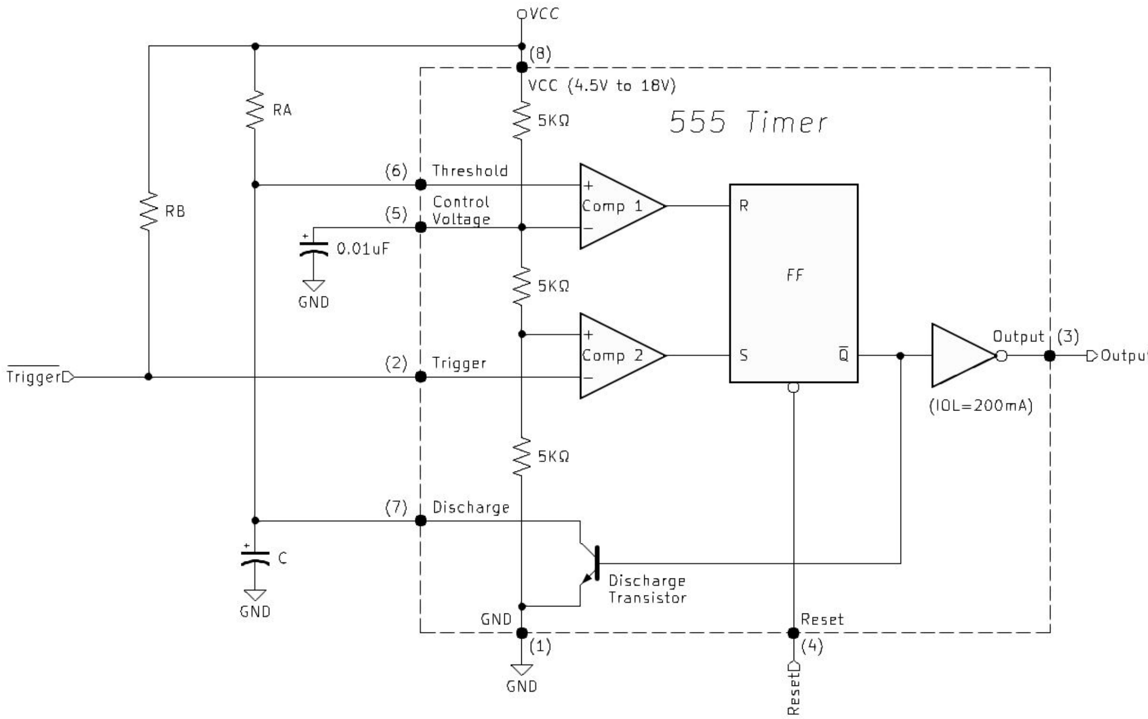
Find: RA, RB.

Solve:

$RA =$ ____________

$RB =$ ____________

4. Analyze a Monostable Multivibrator using a 555 Timer.



Given: $VCC = 15V$, $RA = 32K\Omega$, $RB = 10K\Omega$, $C = 10nF$, (The trigger pulse occurs every $500\mu S$).

Find: PW, DC.

Solve:

$PW =$ ____________

$DC =$ ____________

11c.10 Answers:

1. $R1 = 17.3K\Omega$

   $R2 = 600\Omega$

   $R3 = 28.8K\Omega$

   $R4 = 600\Omega$

   $C1 = 1.9nF$

   $C2 = 2.3nF$

2. $R1 = 1K\Omega$

   $R2 = 27K\Omega$

   $R3 = 1K\Omega$

   $R4 = 54.8K\Omega$

   $C1 = 99.932\mu F$

3. $RA = 47K\Omega$

   $RB = 2K\Omega$

4. $PW = 351.566\mu S$

   $DC = 70.311\%$

Week 12a

Linear Regulators

12a.1 Objectives

Knowledge Objectives

- **Understand Linear Regulator Fundamentals**: Explain the basic principles of linear voltage regulation, including the operation of series and shunt regulators.
- **Types of Linear Regulators**: Identify and describe the different types of linear regulators, such as low-dropout (LDO) regulators, adjustable regulators, and fixed regulators.
- **Key Parameters and Specifications**: Understand and interpret key parameters of linear regulators, including dropout voltage, load regulation, line regulation, quiescent current, and power dissipation.
- **Power Efficiency**: Explain the efficiency of linear regulators and compare it to other types of regulators, such as switching regulators.

Design Objectives

- **Design Simple Linear Regulator Circuits**: Design and implement basic linear regulator circuits for various applications, selecting appropriate components such as transistors, operational amplifiers, and pass elements.
- **Component Selection**: Choose appropriate components for linear regulator circuits based on the required output voltage, current, and power dissipation considerations.
- **Thermal Management**: Design and implement thermal management solutions in linear regulator circuits, including the use of heatsinks and thermal protection circuits.

Troubleshooting Objectives

- **Identify Common Issues**: Identify and diagnose common problems in linear regulator circuits, such as excessive heat generation, instability, and poor load regulation.
- **Testing and Measurement**: Use multimeters, oscilloscopes, and other test equipment to measure voltage regulation, ripple, noise, and response to load changes in linear regulator circuits.

Integrated Objectives

- **Real-world Applications**: Discuss and analyze real-world applications of linear regulators in power supplies, battery-powered devices, and sensitive analog circuits.
- **Comparison with Switching Regulators**: Compare linear regulators with switching regulators, discussing scenarios where one may be preferred over the other based on factors like efficiency, noise, and complexity.
- **Lab Project**: Design, build, and troubleshoot a power supply circuit using a linear regulator, demonstrating a thorough understanding of its design principles and performance characteristics.

12a.2 Introduction

Linear regulators are a fundamental component in power supply design, essential for maintaining a stable and precise output voltage from a varying input voltage source. Unlike switching regulators, linear regulators operate by continuously adjusting the resistance of a pass element, such as a transistor, to drop excess voltage, resulting in a smooth, low-noise output. Although they are less efficient than their switching counterparts due to their inherent power dissipation, linear regulators are prized for their simplicity, low noise, and ability to provide clean power to sensitive analog circuits. Understanding the operation, design, and application of linear regulators is crucial for anyone involved in electronics, particularly in areas where voltage stability and noise minimization are paramount.

12a.3 78XX Linear Regulators

The 78XX series of linear voltage regulators is a widely-used family of fixed-output regulators known for their simplicity, reliability, and ease of use. These regulators provide a stable output voltage of 5V (7805), 9V (7809), 12V (7812), and other standard values, making them ideal for powering a wide range of electronic circuits. The 78XX series operates by maintaining a constant output voltage despite variations in input voltage and load current, making them invaluable in applications where precise voltage regulation is required. With built-in thermal shutdown and short-circuit protection, the 78XX series not only simplifies power supply design but also adds a layer of safety and durability to electronic systems. Understanding the operation and application of the 78XX series is a fundamental step in learning how to design effective and reliable power supplies in electronics.

12a.3.1 78XX Key Features:

See datasheet here: Texas Instruments 7805 TO-220 [24]

- Ease of use
- Available in TO-92, TO-220, and surface mount packages
- Output Current up to 1.5A
- Internal Thermal-Overload Protection
- Internal Short-Circuit Limiting

12a.4 LM317 Linear Regulators

The LM317 is a versatile and widely used adjustable linear voltage regulator that allows for precise control of output voltage, making it a popular choice in a variety of electronic applications. Unlike fixed regulators, the LM317 can be set to provide any output voltage between 1.25V and 37V, simply by adjusting two external resistors. This flexibility, combined with its built-in features like thermal overload protection, current limiting, and safe area protection, makes the LM317 a robust and reliable choice for both hobbyists and professionals. Understanding the LM317 not only equips you with the ability to design custom voltage regulation solutions but also deepens your knowledge of how adjustable regulators function within a circuit.

12a.4.1 LM317 Key Features:

See datasheet here: Onsemi LM317 TO-220 [25]

- Adjustable output from 1.2V to 37V
- Available in TO-92, TO-220, and surface mount packages
- Output Current up to 1.5A
- Internal Thermal-Overload Protection
- Internal Short-Circuit Limiting

12a.5 LM317 Fixed Voltage Regulation

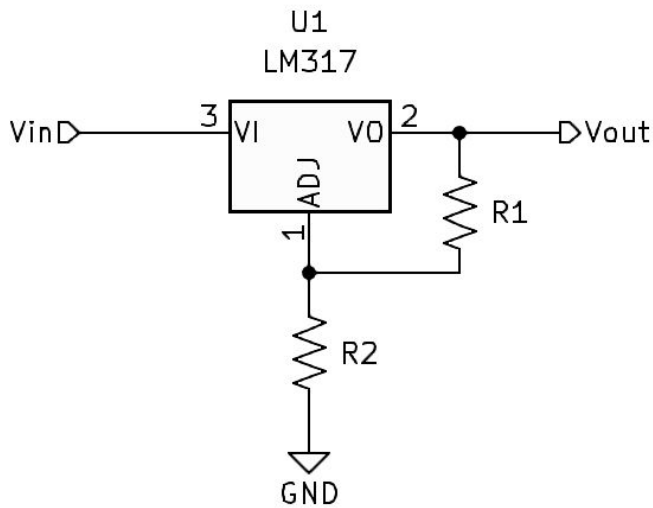


Figure 12a.1: LM317 Fixed Voltage Regulation Circuit

LM317 Design Rules:

- No current into or out of Pin1 ADJ.
- The LM317 will do whatever it can to maintain 1.25V from Pin2 OUT to Pin1 ADJ.
- Make $I_{R1} \approx 5mA$

✓ $R1 \approx \frac{1.25V}{5mA} \approx 250\Omega$

- Choose Vout
- $VR2 = Vout - VR1 = Vout - 1.25V$
- $R2 = \frac{VR2}{IR1} = \frac{Vout-1.25V}{IR1} = \frac{Vout-1.25V}{\frac{1.25V}{R1}} = \frac{R1(Vout-1.25V)}{1.25V}$

✓ $R2 = \frac{R1(Vout-1.25V)}{1.25V}$

12a.6 LM317 Variable Voltage Regulation

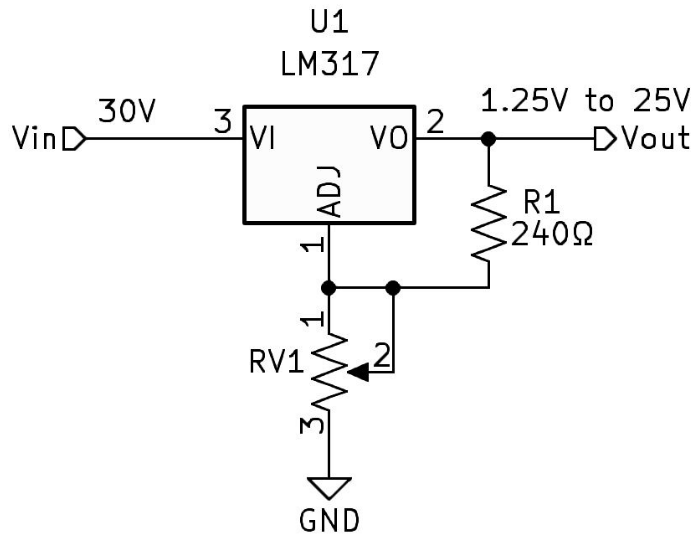


Figure 12a.2: LM317 Variable Voltage Regulation Circuit

Vout=1.25V to 25V

- At the potentiometer RV1 extreme where the center pin (pin2) is tied to ground, the regulated output will be equal to VR1 at 1.25V.
- At the other extreme of the potentiometer RV1 where the center pin (pin2) is tied to pin 1, Vout will equal VR1 plus VRV1 which is the Vout Max value.
- Equation to solve for RV1:
  - $Vout_{max} = IR1 \times RT$
  - $Vout_{max} = \frac{VR1}{R1}(R1 + RV1)$
  - $Vout_{max} = \frac{1.25V}{R1}(R1 + RV1)$
  - $\frac{Vout_{max} \times R1}{1.25V} = (R1 + RV1)$
  - ✓ $RV1 = \frac{Vout_{max} \times R1}{1.25V} - R1$
- Solve for RV1:
  - $RV1 = \frac{Vout_{max} \times R1}{1.25V} - R1$
  - $RV1 = \frac{25V \times 240\Omega}{1.25V} - 240\Omega$
  - $RV1 = 4.56K\Omega$
  - ✓ Tested circuit with a standard value $RV1 = 5K\Omega$, measured 1.2V to 28.4V.

12a.7 LM317 Fixed Current Regulation

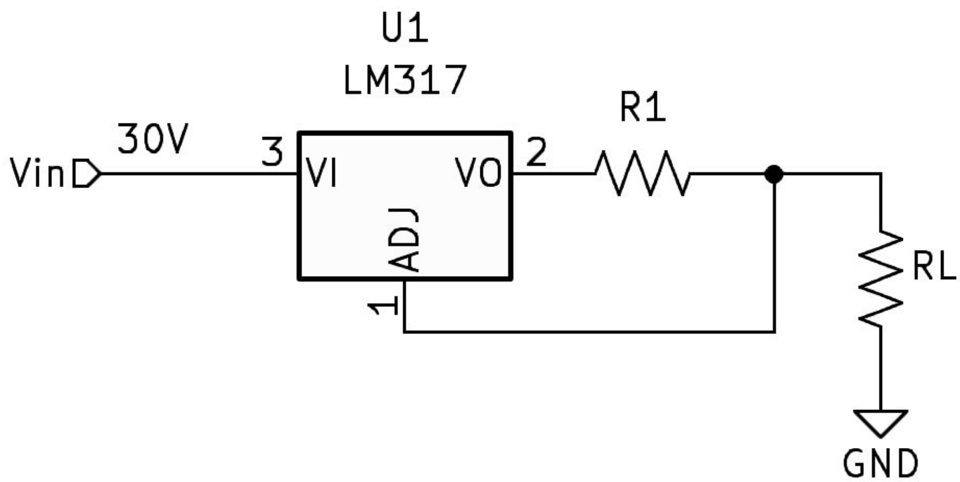


Figure 12a.3: LM317 Fixed Current Regulation Circuit

LM317 Fixed Current Limiting Calculations:

- $Iout_{max} = \frac{VR1}{R1} = \frac{1.25V}{R1}$
- $R1 = \frac{VR1}{Iout_{max}} = \frac{1.25V}{Iout_{max}}$
- $P_{R1} = IR1 \times VR1 = Iout_{max} \times 1.25V$

12a.8 LM317 Variable Current Regulation

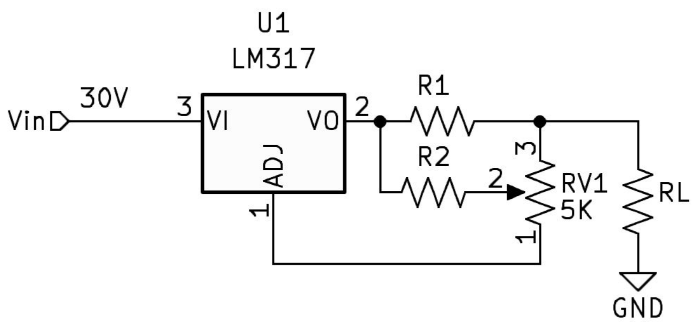


Figure 12a.4: LM317 Variable Current Regulation Circuit

R1 Calculations:

To solve for R1, RV1 will be in the extreme position where its pins 2 and 3 are connected. This position represents $Imin$. To better analyze the circuit an $Imin$ circuit redraw Figure 12a.5 is provided.

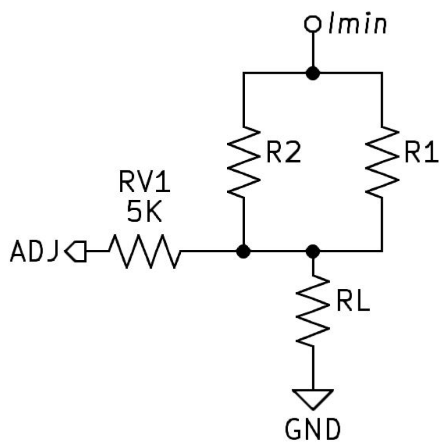


Figure 12a.5: LM317 $Imin$ Circuit Redraw

- There is no significant current flow into or out of the ADJ pin 1 of the LM317. Therefore, there will be no voltage across RV1.
- R1 is now in parallel with R2 and begin limiting the current when $VR2$ reaches 1.25V.
- $R1 = \frac{VR2}{I_{min}} = \frac{1.25V}{I_{min}}$

✓ $R1 = \frac{1.25V}{I_{min}}$

R2 Calculations:

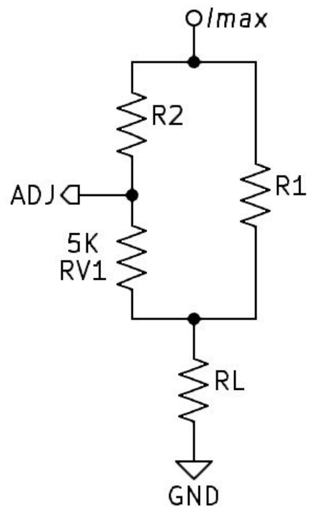


Figure 12a.6: LM317 *Imax* Circuit Redraw

- The LM317 *Imax* circuit redraw Figure 12a.6. shows RV1 in the extreme position where pins 1 and 2 are connected.
- The current through the R2-RV1 branch is insignificant compared to the IR1 current making IR1 nearly equivalent to Imax.
- $IR1 = Imax$
- $VR1 = IR1 \times R1 = Imax \times R1$
- $VR2 = 1.25V$
- $V_{RV1} = VR1 - VR2 = VR1 - 1.25V = (Imax \times R1) - 1.25V$
- $I_{RV1} = \frac{V_{RV1}}{RV1} = \frac{V_{RV1}}{5K\Omega} = \frac{(Imax \times R1) - 1.25V}{5K\Omega} = \frac{(Imax \times R1) - 1.25V}{RV1}$
- $R2 = \frac{VR2}{I_{RV1}} = \frac{1.25}{\frac{(Imax \times R1) - 1.25V}{RV1}} = \frac{RV1 \times 1.25V}{(Imax \times R1) - 1.25V} = \frac{RV1 \times 1.25V}{(Imax \times \frac{1.25}{Imin}) - 1.25V}$

✓ $R2 == \frac{RV1 \times 1.25V}{(\frac{Imax \times 1.25}{Imin}) - 1.25V}$

12a.9 LM317 Variable Voltage with Variable Current

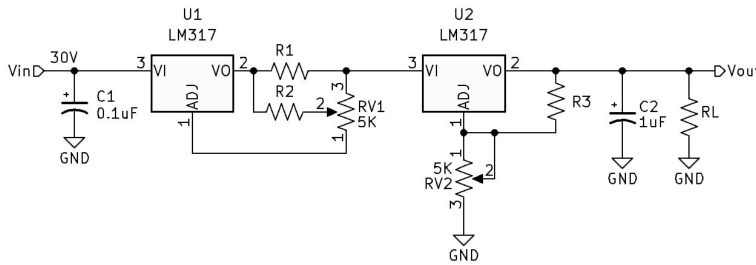


Figure 12a.7: LM317 Regulated Variable Voltage with Variable Current Circuit

12a.10 LM317 Variable Voltage with Variable Current and Over-Voltage

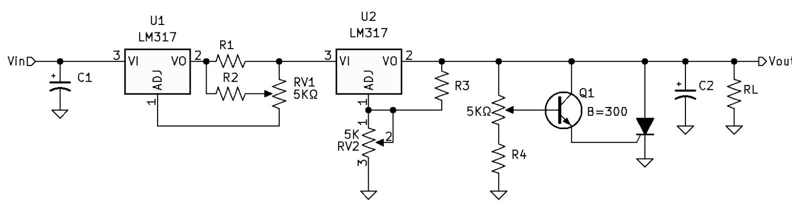


Figure 12a.8: LM317 Regulated Variable Voltage with Variable Current Circuit and Over-Voltage Protection

12a.11 Split Rail LM317 & LM337

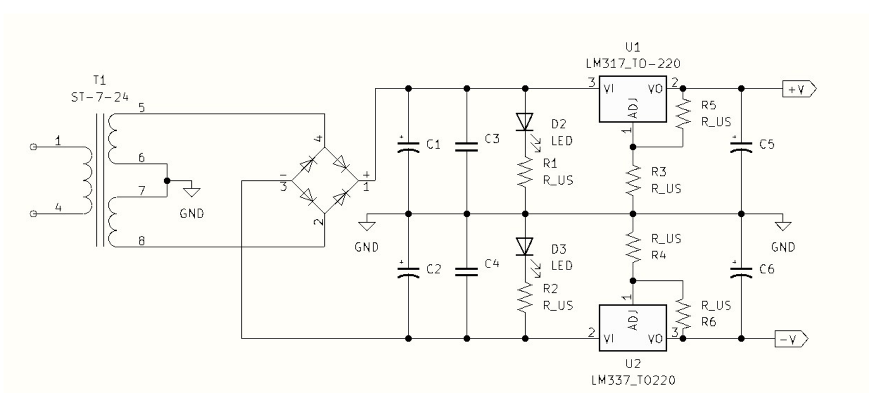


Figure 12a.9: Split Rail Power Supply with LM317 & LM337

12a.12 Heatsink Testing using the LM317

Refer to section 5c on page 145 to review Heatsink Calculations. Taking advantage of the LM317's built-in temperature protection capability can provide relatively accurate testing and verification of our previously predicted Heatsink calculations. For example, if you have a heatsink with a predicted maximum power dissipation of 10 watts, design a circuit using the LM317 that will require more than 10 watts of power dissipation, maybe 15 or 20 watts (too much may cause the current limiting to interfere with the temp limiting). Connect the LM317 to the heatsink using thermal paste and test. Initially, the circuit will operate as calculated. Once the heatsink heats up and begins to saturate the LM317 will start to overheat which will cause it to decrease the output current, decreasing the power across the LM317. The lowering voltage at the output will eventually stabilize, allowing the operator to measure the true power dissipation of the heatsink.

12a.13 Practice Problems

1. According to the Datasheet, what is the absolute maximum input voltage for the 7805? See datasheet here: Texas Instruments 7805 TO-220 [24]

   7805 $Vin_{AbsoluteMax} =$ ____________

2. According to the Datasheet, what is the recommended minimum and maximum input voltages for the 7805? See datasheet here: Texas Instruments 7805 TO-220 [24]

   7805 $Vin_{RecommendedMin} =$ ____________

   7805 $Vin_{RecommendedMax} =$ ____________

3. According to the Datasheet, what is the maximum output current for the 7805? See datasheet here: Texas Instruments 7805 TO-220 [24]

   7805 $Iout_{Max} =$ ____________

4. According to the Datasheet, what is the maximum differential voltage for the LM317? See datasheet here: Onsemi LM317 TO-220 [25]

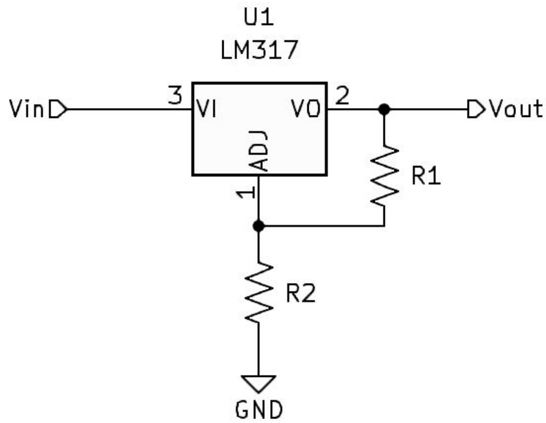
   LM317 $Vdiff_{Max} =$ ____________

5. According to the Datasheet, what is the Thermal Resistance Junction to Case, T Package for the LM317? See datasheet here: Onsemi LM317 TO-220 [25]

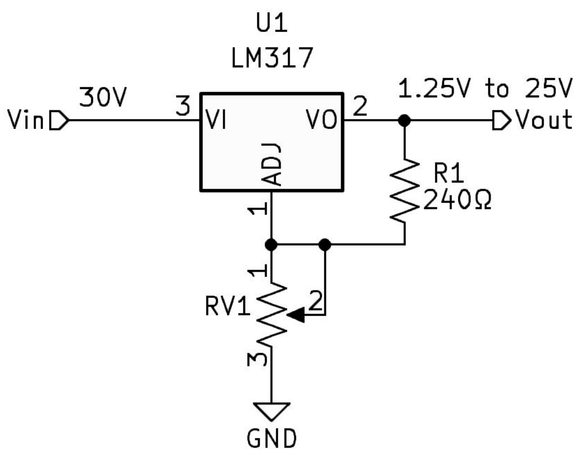
   LM317 $R_{\theta JC} =$ ____________

6. Design a 12V Fixed Regulator using an LM317. Make $R1 = 240\Omega$.

   $R2 =$ ____________



7. Design a Variable Voltage Regulator with an output capable of 1.25V to 25V using an LM317. Make $R1 = 240\Omega$.

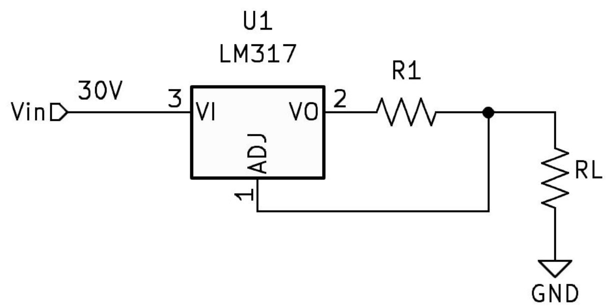


$RV1 =$ ____________

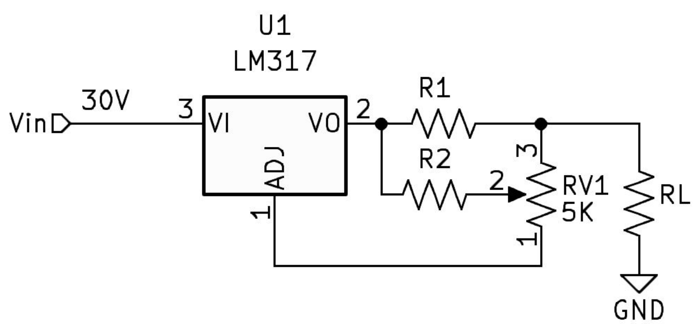
8. Design a Fixed Current Limiting Circuit to provide a maximum regulated current of 0.5 amps.

$R1 =$ ____________

$P_{R1} =$ ____________



9. Design a Variable Current Limiting Circuit capable of providing a regulated 0.2 amps to 1.0 amps. Make RV1 equal to $5K\Omega$.



$R1 =$ ____________

$R2 =$ ____________

$P_{R1} =$ ____________

12a.14 Practice Problem Answers

1. 7805 $Vin_{AbsoluteMax} = 35V$
2. 7805 $Vin_{RecommendedMin} = 7V$, 7805 $Vin_{RecommendedMax} = 25V$
3. 7805 $Iout_{max} = 1.5A$
4. LM317 $Vdiff_{Max} = 40V$
5. LM317 $R_{\theta JC} = 5°C/W$
6. $R2 = 2.064K\Omega$
7. $RV1 = 4.56K\Omega$
8. $R1 = 2.5\Omega,\ P_{R1} = 625mW$
9. $R1 = 6.25\Omega,\ R2 = 1.25K\Omega,\ P_{R1} = 6.25W$

Week 12b

Switching Power Supplies

12b.1 Objectives

Knowledge Objectives:

1. **Understand Switching Power Supply Fundamentals**: Explain the basic principles of switch-mode power supplies (SMPS), including the role of switching elements, energy storage components, and feedback control.

2. **Types of Switching Power Supplies**: Identify and describe the different types of SMPS topologies, such as buck, boost, buck-boost, flyback, and forward converters.

3. **Key Parameters and Specifications**: Understand and interpret key parameters and specifications of switching power supplies, including efficiency, ripple voltage, switching frequency, and power factor.

4. **Operation Modes**: Differentiate between continuous and discontinuous conduction modes and understand their impact on the performance and design of switching power supplies.

Design Objectives:

1. **Design Simple Switching Power Supply Circuits**: Design and implement basic SMPS circuits using various topologies, selecting appropriate components such as inductors, capacitors, diodes, and MOSFETs.

2. **Component Selection**: Choose appropriate components for SMPS circuits based on factors such as voltage, current, power rating, and thermal considerations.

3. **Feedback and Control**: Design and implement feedback control loops for switching power supplies to regulate output voltage and ensure stability.

Troubleshooting Objectives:

1. **Identify Common Issues**: Identify and diagnose common problems in SMPS circuits, such as excessive noise, instability, electromagnetic interference (EMI), and overheating.

2. **Testing and Measurement**: Use oscilloscopes, multimeters, and other test equipment to measure key performance metrics, including output voltage, efficiency, ripple, and switching waveforms.

Integrated Objectives:

1. **Real-world Applications**: Discuss and analyze real-world applications of switching power supplies in various devices and systems, including consumer electronics, industrial equipment, and renewable energy systems.

2. **Comparison with Linear Power Supplies**: Compare switching power supplies with linear power supplies, discussing scenarios where one may be preferred over the other based on factors like efficiency, size, cost, and noise.

3. **Lab Project**: Design, build, and troubleshoot a complete switching power supply circuit, demonstrating a thorough understanding of its design principles, control methods, and performance characteristics.

12b.2 Introduction

Switching power supplies, also known as switch-mode power supplies (SMPS), are highly efficient power conversion systems that are widely used in modern electronics due to their ability to convert electrical energy with minimal losses. Unlike linear regulators, which dissipate excess energy as heat, switching power supplies use high-speed electronic switches to rapidly alternate the input voltage, storing energy in inductors or capacitors and then releasing it at the desired output voltage. This approach allows switching power supplies to achieve much higher efficiency, making them ideal for applications where power conservation, compact size, and heat management are critical. Understanding the principles, design, and operation of switching power supplies is essential for anyone involved in electronics, as they are a cornerstone of power management in everything from consumer electronics to industrial systems.

12b.3 SMPS Topologies

12b.3.1 BUCK

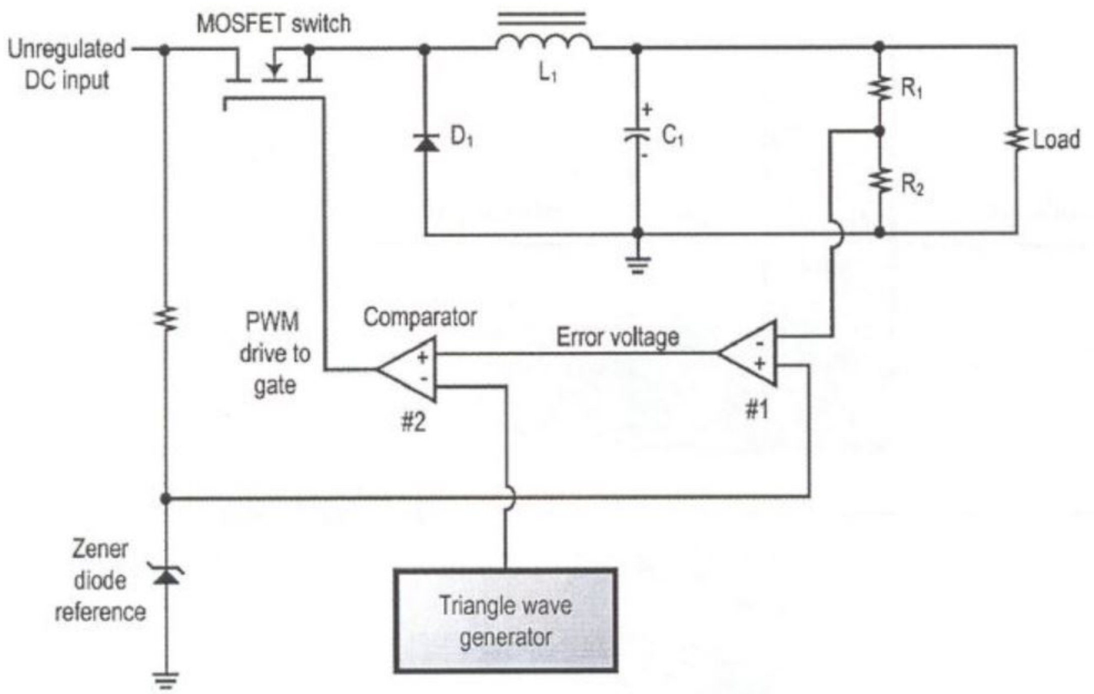


Figure 12b.1: BUCK SMPS with Regulation

The Buck Switching Mode Power Supply (SMPS) topology is a fundamental and widely used method for step-down voltage conversion in electronic circuits. It efficiently reduces a higher input voltage to a lower output voltage by rapidly switching a transistor on and off, controlling the flow of current through an inductor and storing energy in the magnetic field. The energy is then released in a controlled manner, providing a stable and lower output voltage. The Buck converter's high efficiency, typically above 90%, makes it ideal for applications where power conservation and heat management are crucial, such as in battery-powered devices and low-voltage digital circuits.

12b.3.2 Boost

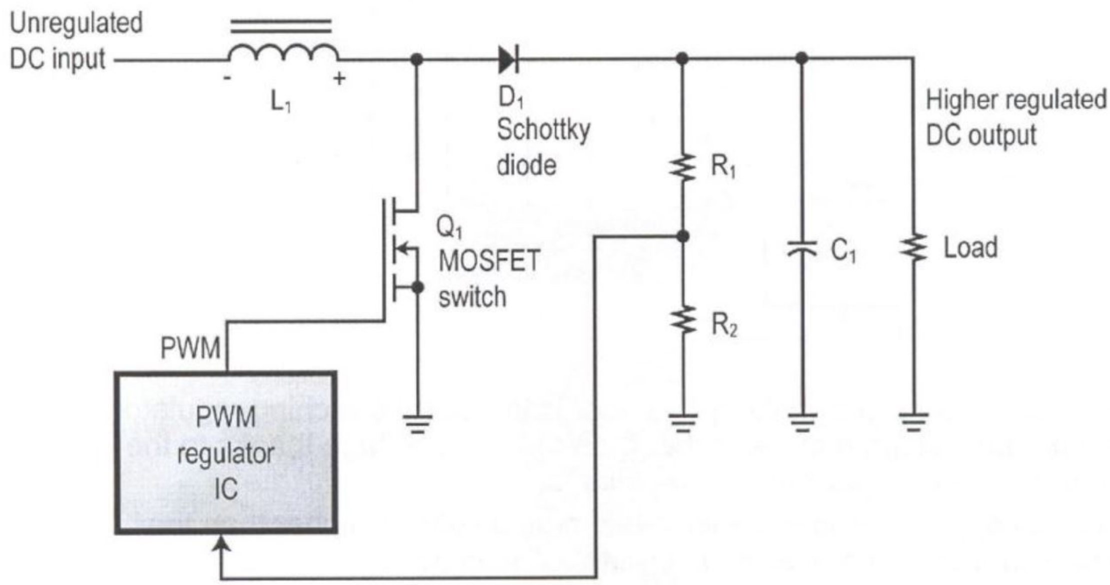


Figure 12b.2: Boost SMPS with Regulation

The Boost Switch Mode Power Supply (SMPS) topology is a crucial method for stepping up a lower input voltage to a higher output voltage in electronic systems. It operates by using a switching transistor and an inductor to temporarily store energy from the input source, then releasing it at a higher voltage when the switch is turned off. This efficient energy conversion process allows the Boost converter to increase voltage while maintaining relatively high efficiency, making it ideal for applications where a higher voltage is needed from a low-voltage power source, such as in battery-powered devices, renewable energy systems, and automotive electronics.

12b.3.3 Buck-Boost Inverter

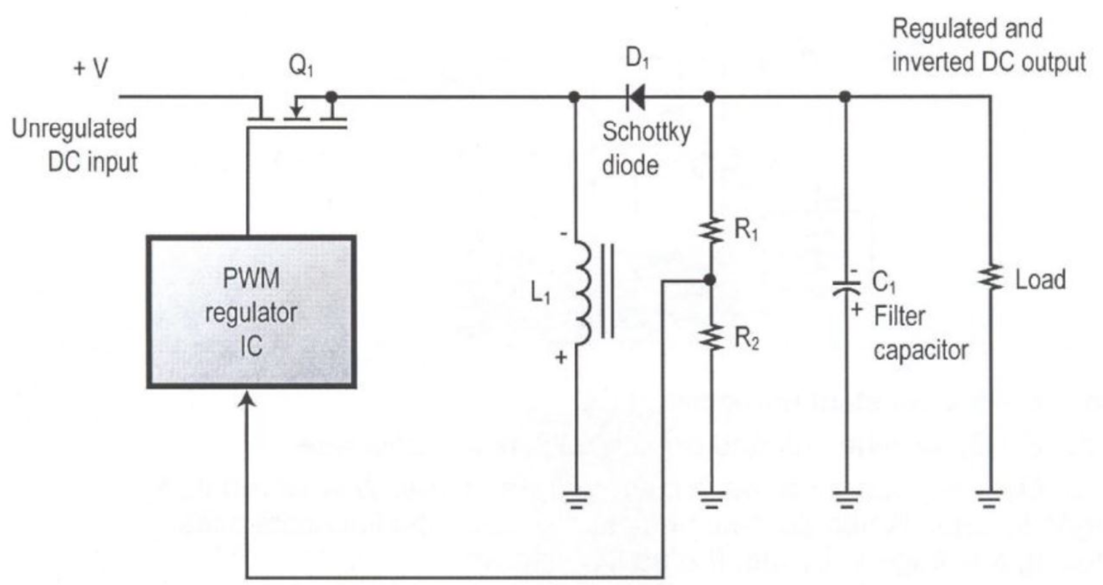


Figure 12b.3: Buck-Boost Inverter SMPS with Regulation

The Buck-Boost Inverter Switch Mode Power Supply (SMPS) is a specialized topology designed to efficiently convert a DC input voltage into an inverted output voltage, meaning the polarity of the output is opposite to that of the input. This topology combines the features of both buck and boost converters, allowing it to either step up or step down the input voltage while also inverting it. This makes the Buck-Boost Inverter particularly valuable in applications where a specific negative voltage is required, such as in bipolar power supplies, operational amplifier circuits, and certain types of sensor systems. By leveraging the advantages of high efficiency and flexibility, the Buck-Boost Inverter provides a reliable solution for powering circuits that demand a stable inverted voltage, making it an essential concept in advanced power supply design.

12b.3.4 Additional Information on SMPS Topologies

- Video - Texas Instruments Power Topologies Overview [26]
- Texas Instruments Power Topologies Handbook [27]
- Texas Instruments Power Topologies Quick Reference Guide [28]

12b.4 Discrete BUCK SMPS

General Design Considerations and Formulas:

- Design at Maximum Current
- Duty Cycle $\approx \frac{Vout}{Vin} \approx 50\%$
- Switching Frequency $(50Khz\ to\ 500Khz)$
- $V_{ind} = -L\frac{di}{dt}$
- $I_C = C\frac{dv}{dt}$

Review the following data sheets:

- BS170 Data Sheet[29]
- IRF9Z24 Data Sheet[30]

12b.4.1 Verify Switch Operation

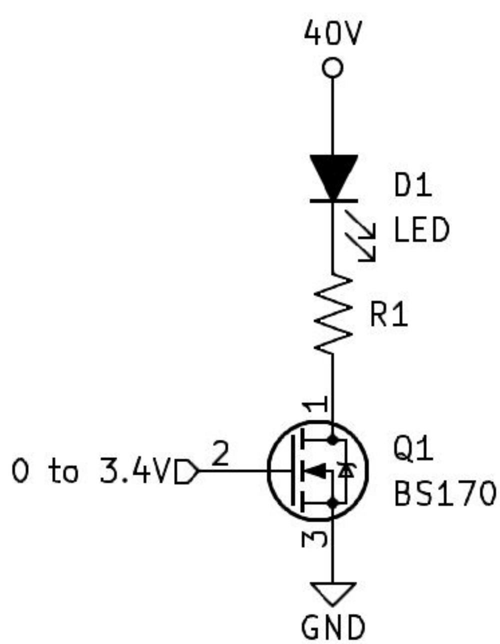


Figure 12b.4: BS170 Test Circuit

- Design a circuit that will test and verify the switching operation of a BS170.

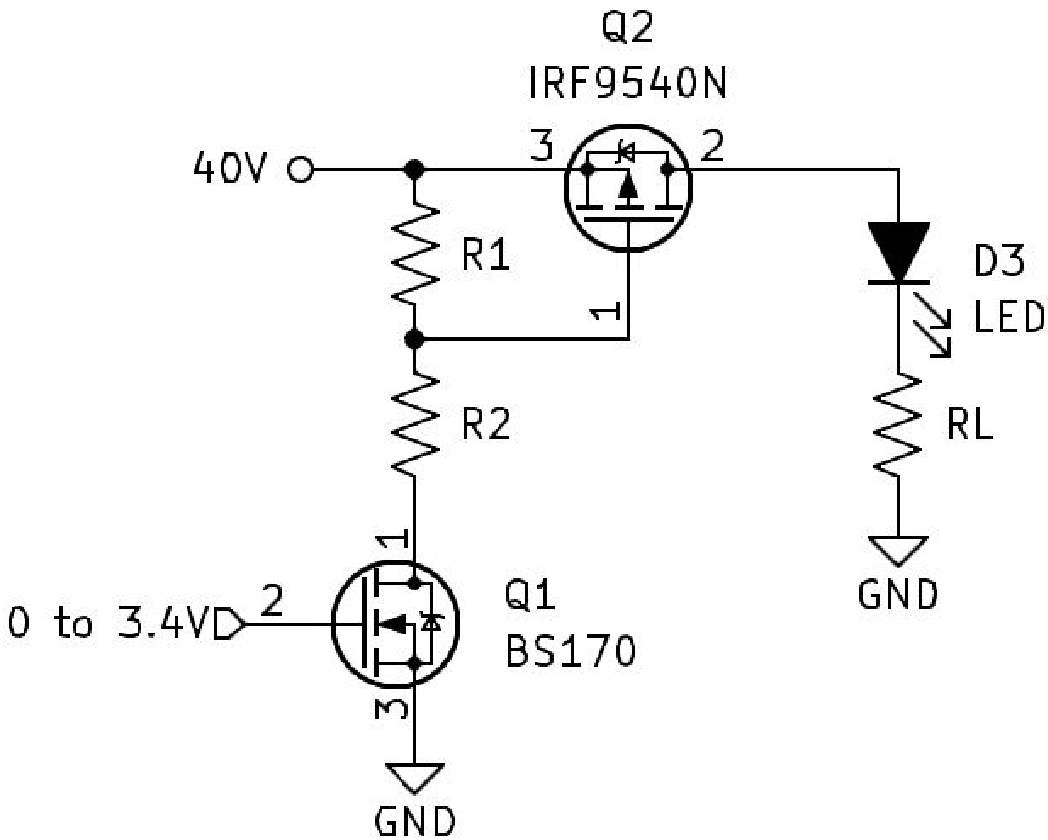


Figure 12b.5: BS170 and IRF9540N Test Circuit

- The TO-220 IRF9540N can handle more power than the TO-92 BS170. Design a circuit that will test and verify the switching operation of the IRF9540N.

- (R1 and R2) When Q1 is on, R1 and R2 act like a voltage divider. VR1 provides the necessary VGS to turn on Q2. What is the Absolute Maximum VGS according to the data sheet for the IRF9540N. There is a inverse relationship between RDS (resistance drain to source) and VGS, meaning more VGS will decrease RDS. Reducing RDS will improve the efficiency of the SMPS however, to prevent damage VGS must be less than VGS_{Max}.

- $RL = \frac{40V-3V}{5mA}$

- Use a Function Generator, 555 Timer, or Astable BJT Multivibrator to clock Q1 slow enough to see the LED blink.

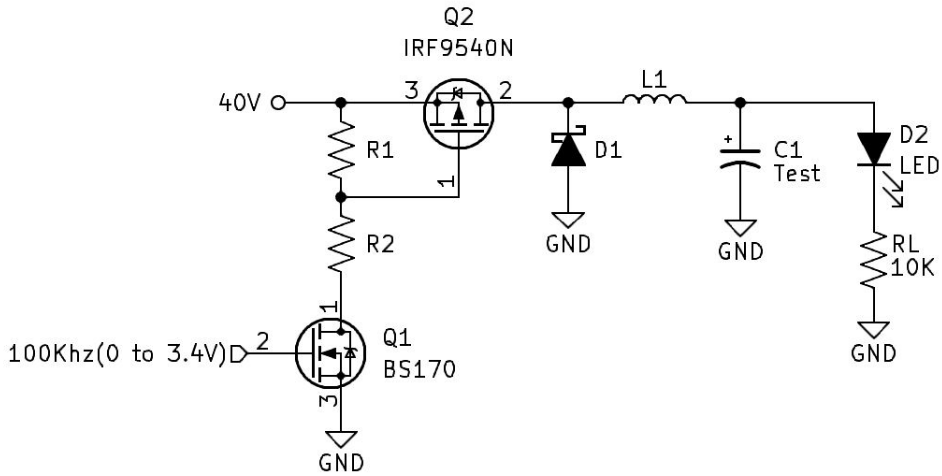


Figure 12b.6: Discrete Buck Test Circuit 1

Buck SMPS Design:

- Using a $10K\Omega$ load the basic Buck SMPS can be tested in a low load scenario.
- (R1 and R2) When Q1 is on, R1 and R2 act like a voltage divider. VR1 provides the necessary VGS to turn on Q2. What is the Absolute Maximum VGS according to the data sheet for the IRF9540N. There is a inverse relationship between RDS (resistance drain to source) and VGS, meaning more VGS will decrease RDS. Reducing RDS will improve the efficiency of the SMPS however, to prevent damage VGS must be less than VGS_{Max}.
- Design for a Duty Cycle $\approx \frac{Vout}{Vin} \approx 50\%$
  - $Vin = 40V$
  - $Vout = 25V$
  - $\frac{Vout}{Vin} = \frac{25}{40} = 62.5\% \approx 50\%$
- Design for a Maximum Current of 1 amp.
- Pick a frequency (50Khs to 500Khs).
  - 100Khz
- $V_{ind} = L\frac{di}{dt}$
  - $25V = L\frac{ripplecurrent}{PW}$
  - Determine the acceptable ripple current:
    - ✻ 1% ripple = 1A to 0.99A, di = 0.01A

        - ✻ 5% ripple = 1A to 0.95A, di = 0.05A
        - ✻ 10% ripple = 1A to 0.90A, di = 0.10A
    - ○ $dt = PW = DC\frac{1}{f} = 0.5\frac{1}{100Khz} = 5\mu Sec$

- Solve for L at various ripple currents:
    - ○ At 1%:
        - ✻ $25V = L\frac{0.01A}{5\mu S}$
        - ✻ $L = 25V\frac{5\mu S}{0.01A}$
        - ✻ $L = 12.5mH$
    - ○ At 5%:
        - ✻ $25V = L\frac{0.05A}{5\mu S}$
        - ✻ $L = 25V\frac{5\mu S}{0.05A}$
        - ✻ $L = 2.5mH$
    - ○ At 10%:
        - ✻ $25V = L\frac{0.1A}{5\mu S}$
        - ✻ $L = 25V\frac{5\mu S}{0.1A}$
        - ✻ $L = 1.25mH$

- Choose a standard value inductor for an acceptable ripple:
    - ○ $10mH$

- Solve for C using $IC = C\frac{dv}{dt}$:
    - ○ Find di for the 10mH inductor:
        - ✻ $Vind = L\frac{di}{dt}$
        - ✻ $25V = 10mH\frac{di}{5\mu S}$
        - ✻ $di = 12.5mA$
        - ✻ $I = 1A$, $(I_{max} = 1.00625A$ to $_{min} = 993.75mA)$
    - ○ $RL = \frac{V}{I} = \frac{25V}{1A} = 25\Omega$
    - ○ $V_{max} = I_{max} \times RL = 1.00625mA \times 25\Omega = 25.1563V$
    - ○ $V_{min} = I_{min} \times RL = 993.75mA \times 25\Omega = 24.844V$
    - ○ $dv = V_{max} - V_{min} = 25.1563V - 24.844V = 312.5mV$
    - ○ $dt = PW = DC\frac{1}{f} = 0.5\frac{1}{100Khz} = 5\mu Sec$
    - ○ $IC = C\frac{dv}{dt}$
        - ✻ $1A = C\frac{312.5mV}{5\mu S}$
        - ✻ $C = 1A\frac{5\mu S}{312.5mV}$
        - ✓ $C \geq 16\mu F$

- Solve for C using $VC = Vfin - (Vfin - Vin)e^{\frac{-t}{RC}}$:
  - $24.844V = 0 - (0 - 25.1563V)e^{\frac{-5\mu S}{25\Omega C}}$
  - $\frac{24.844V}{25.1563V} = e^{\frac{-5\mu S}{25\Omega C}}$
  - $LN\frac{24.844V}{25.1563V} = \frac{-5\mu S}{25\Omega C}$
  - $C = \frac{-5\mu S}{25\Omega \times (LN\frac{24.844V}{25.1563V})}$
  - ✓ $C \geq 16\mu F$

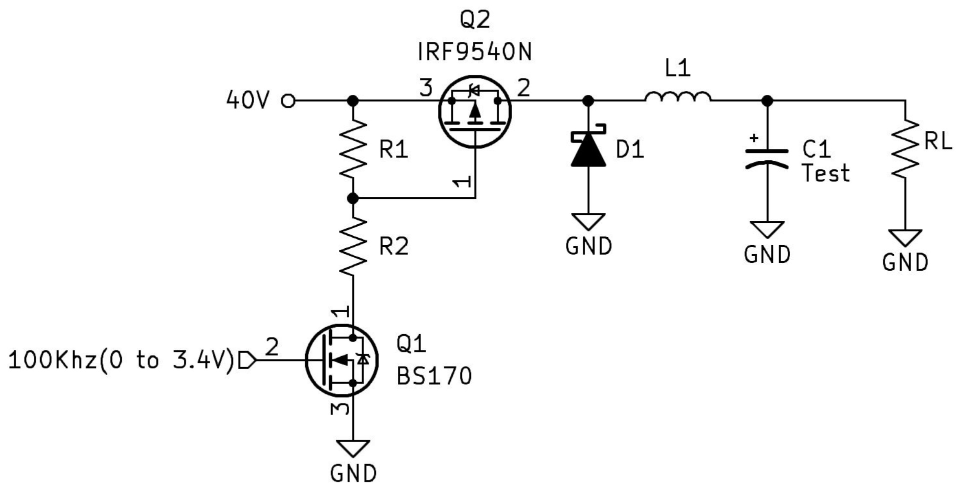


Figure 12b.7: Discrete Buck Test Circuit 2

Table 12b.1: Test Data Discrete BUCK SMPS

Load	Duty Cycle	C1 Capacitance	L1 Inductance	Vout
$10K\Omega$	20%	$10\mu F$	$10mH$	6.8V
$10K\Omega$	20%	$47\mu F$	$10mH$	6.8V
$10K\Omega$	20%	$100\mu F$	$10mH$	6.8V
$10K\Omega$	30%	$10\mu F$	$10mH$	9.9V
$10K\Omega$	30%	$47\mu F$	$10mH$	9.8V
$10K\Omega$	30%	$100\mu F$	$10mH$	9.9V
$10K\Omega$	40%	$10\mu F$	$10mH$	17.7V
$10K\Omega$	40%	$47\mu F$	$10mH$	13.3V
$10K\Omega$	40%	$100\mu F$	$10mH$	13.3V
$10K\Omega$	50%	$10\mu F$	$10mH$	20.1V
$10K\Omega$	50%	$47\mu F$	$10mH$	20.0V
$10K\Omega$	50%	$100\mu F$	$10mH$	20.1V
$10K\Omega$	60%	$10\mu F$	$10mH$	37.5V
$10K\Omega$	60%	$47\mu F$	$10mH$	37.5V
$10K\Omega$	60%	$100\mu F$	$10mH$	37.5V

Table 12b.2: Test Data Discrete BUCK SMPS

Load	Duty Cycle	C1 Capacitance	L1 Inductance	Vout
$4.7K\Omega$	40%	$10\mu F$	$10mH$	10.5V
$4.7K\Omega$	50%	$10\mu F$	$10mH$	14.3V
$4.7K\Omega$	60%	$10\mu F$	$10mH$	29.6V
$2.2K\Omega$	40%	$10\mu F$	$10mH$	9.2V
$2.2K\Omega$	50%	$10\mu F$	$10mH$	12.9V
$2.2K\Omega$	60%	$10\mu F$	$10mH$	19.6V
$2.2K\Omega$	63%	$10\mu F$	$10mH$	27.1V
$1.1K\Omega$	40%	$10\mu F$	$10mH$	8.3V
$1.1K\Omega$	50%	$10\mu F$	$10mH$	11.4V
$1.1K\Omega$	60%	$10\mu F$	$10mH$	15.7V
$1.1K\Omega$	63%	$10\mu F$	$10mH$	18.4V
$1.1K\Omega$	66%	$10\mu F$	$10mH$	26.5V
500Ω 10W	40%	$10\mu F$	$10mH$	9.7V
500Ω 10W	50%	$10\mu F$	$10mH$	13.3V
500Ω 10W	60%	$10\mu F$	$10mH$	24.7V
500Ω 10W	63%	$10\mu F$	$10mH$	32.8V
300Ω 10W	40%	$10\mu F$	$10mH$	9.1V
300Ω 10W	50%	$10\mu F$	$10mH$	12.6V
300Ω 10W	60%	$10\mu F$	$10mH$	18.9V
300Ω 10W	63%	$10\mu F$	$10mH$	25.6V
100Ω 10W	40%	$10\mu F$	$10mH$	7.3V
100Ω 10W	50%	$10\mu F$	$10mH$	10.2V
100Ω 10W	60%	$10\mu F$	$10mH$	14V
100Ω 10W	63%	$10\mu F$	$10mH$	16.5V
100Ω 10W	65%	$10\mu F$	$10mH$	20V
100Ω 10W	67%	$10\mu F$	$10mH$	24.8V
100Ω 10W	68%	$10\mu F$	$10mH$	27.1V w/ripple
100Ω 10W	68%	$100\mu F$	$10mH$	27.1V improved

Table 12b.3: Test Data Discrete BUCK SMPS

Load	Duty Cycle	C1	L1	Vout	Ripple
65Ω 50W	40%	100μF	10mH	7.0V	minimal
65Ω 50W	50%	100μF	10mH	10.1V	minimal
65Ω 50W	60%	100μF	10mH	14.6V	minimal
65Ω 50W	63%	100μF	10mH	17.6V	minimal
65Ω 50W	65%	100μF	10mH	21.1V	minimal
65Ω 50W	67%	100μF	10mH	24.8V	minimal
65Ω 50W	68%	100μF	10mH	26.6V	300mVpp
25Ω 50W	40%	100μF	10mH	7.1V	minimal
25Ω 50W	50%	100μF	10mH	10.7V	minimal
25Ω 50W	60%	100μF	10mH	16.2V	minimal
25Ω 50W	63%	100μF	10mH	20.3V	minimal
25Ω 50W	65%	100μF	10mH	22.2V	minimal
25Ω 50W	67%	100μF	10mH	24.5V	minimal
25Ω 50W	68%	100μF	10mH	25.9V	800mVpp
25Ω 50W	68%	220μF	10mH	25.9V	144mVpp

Calculating Ripple Factor:

The circuit was measured at 25VDC with a 25Ω load using a variety of capacitor values. The ripple peak-to-peak voltage was measured which appeared to be mostly sinusoidal.

- Ripple Factor $= \frac{V_{RMS}}{V_{DC}} = \frac{\frac{V_{RipplePP}}{2\sqrt{2}}}{25V}$

Table 12b.4: Test Data Discrete BUCK SMPS

Load	Vout DC	C1	Ripple	R-Factor	% Ripple
25Ω 50W	25V	$10\mu F$	$1.84vpp$	0.026	2.6%
25Ω 50W	25V	$47\mu F$	$960mvpp$	0.0135765	1.36%
25Ω 50W	25V	$100\mu F$	$600mvpp$	0.0085	0.85%
25Ω 50W	25V	$220\mu F$	$216mvpp$	0.0031	0.31%

12b.5 Regulated Discrete BUCK SMPS

12b.5.1 TL494 IC

The TL494 is a versatile and widely-used Pulse Width Modulation (PWM) controller IC, designed for managing the operation of switching mode power supplies (SMPS). It provides a complete control solution for various SMPS topologies, including buck, boost, and buck-boost converters, by generating precise PWM signals that regulate the switching of power transistors. The TL494 includes essential features like error amplifiers, a voltage reference, a dead-time control, and an on-chip oscillator, making it a powerful yet compact solution for designing efficient and stable power supplies. Its flexibility and ease of integration have made the TL494 a popular choice in both commercial and DIY power supply projects, allowing designers to implement reliable voltage and current regulation with minimal external components. Understanding the TL494 is crucial for anyone looking to design or optimize SMPS circuits.

12b.5.2 TL494 Data Sheet

- TL494 Data Sheet [31]

7.3 Recommended Operating Conditions

			MIN	MAX	UNIT
V_{CC}	Supply voltage		7	40	V
V_I	Amplifier input voltage		–0.3	$V_{CC} - 2$	V
V_O	Collector output voltage			40	V
	Collector output current (each transistor)			200	mA
	Current into feedback terminal			0.3	mA
f_{OSC}	Oscillator frequency		1	300	kHz
C_T	Timing capacitor		0.47	10000	nF
R_T	Timing resistor		1.8	500	kΩ
T_A	Operating free-air temperature	TL494C	0	70	°C
		TL494I	–40	85	

Figure 12b.8: TL494 Recommended Opperating Conditions

12b.5.3 TL494 Application Test Circuit Analysis:

Oscillator Frequency:

- $f_{osc} = \frac{1}{*R_T \times *C_T} =$ (1Khz to 300Khz)
  - $500K\Omega \geq R_T \geq 1.8K\Omega$
  - $10\mu F \geq C_T \geq 0.47nF$

10.1 Application Information

The following design example uses the TL494 to create a 5-V/10-A power supply. This application was taken from application note SLVA001.

10.2 Typical Application

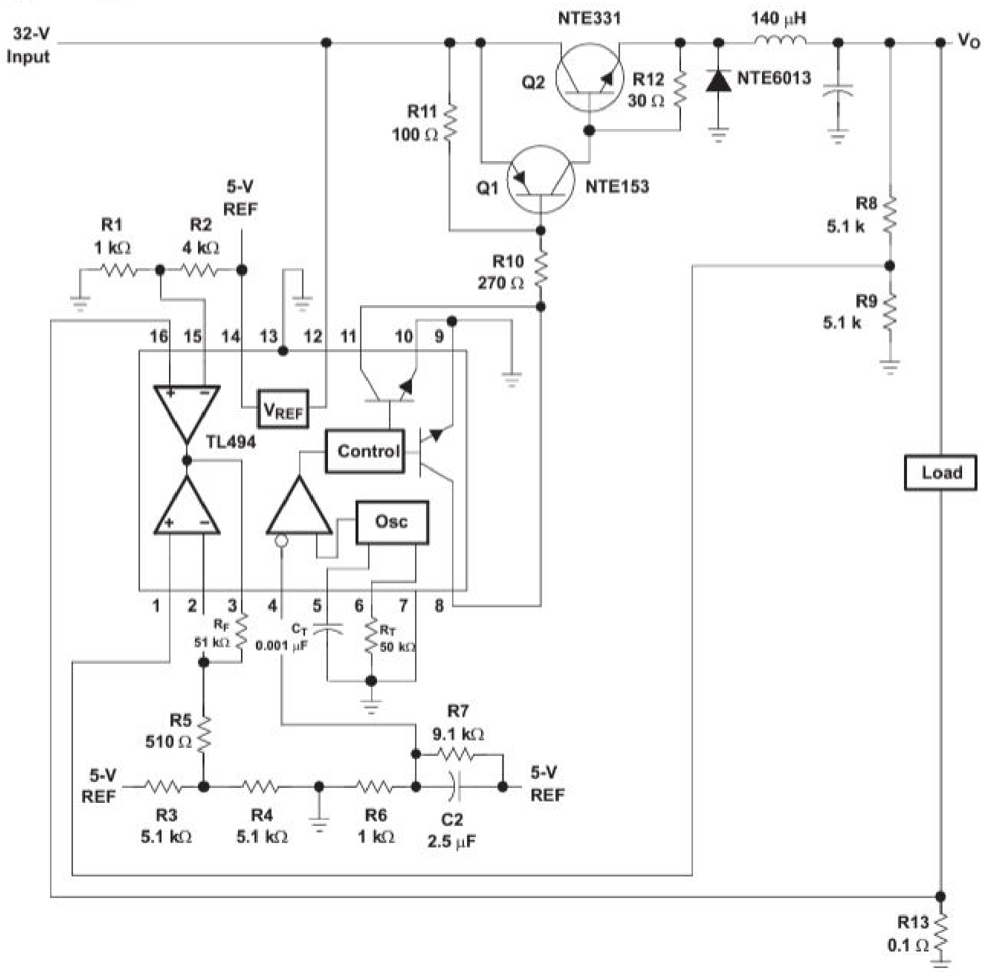


Figure 12b.9: TL494 Application Test Circuit

Current Limiting:

The voltage divider resistors R1 and R2 provide a reference voltage of 1V for pin 15 of the internal op-amp of the TL494. Pin 16 monitors the voltage across R13. R13 is connected to the reference and the load, when the output current reaches the point where R13's voltage is 1V, the TL494 error amplifier will internally begin to limit the output current see figure 12b.10.

- $I_{max} = \frac{1V}{R13} = \frac{1V}{1\Omega} = 1A$

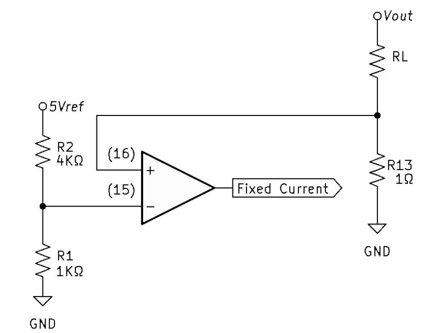


Figure 12b.10: TL494 Fixed Current Limiting

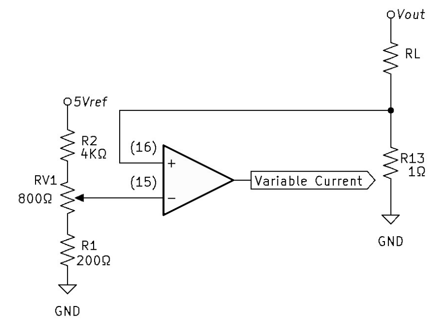


Figure 12b.11: TL494 Variable Current Limiting

Voltage Regulation:

See Figure 12b.12. Voltage divider R3 and R4 are used to set the reference voltage for the error amplifier. According to the data sheet, R5 and RF are used to stabilize the error amplifier. The feedback voltage from the output to pin 1 of the error amplifier will cause the PWM signal to reduce Pulse Width when the voltage is higher than the reference voltage and will increase Pulse Width when the voltage is lower than the reference voltage. Essentially, the error amp will try to keep the two inputs equal by adjusting the Duty Cycle of the PWM.

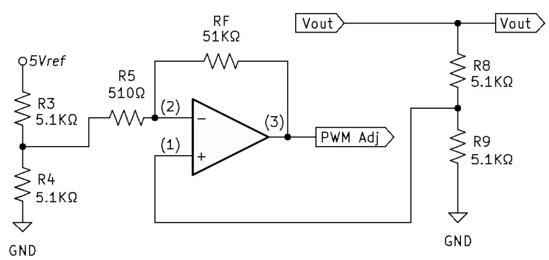


Figure 12b.12: TL494 5V Regulation

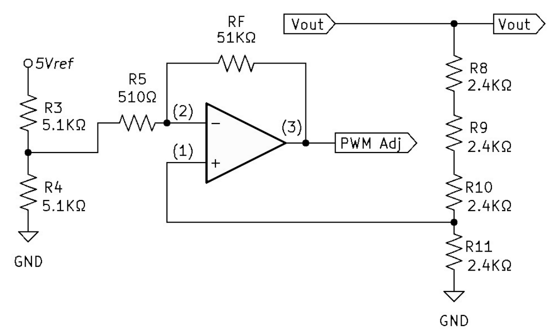


Figure 12b.13: TL494 Fixed Voltage

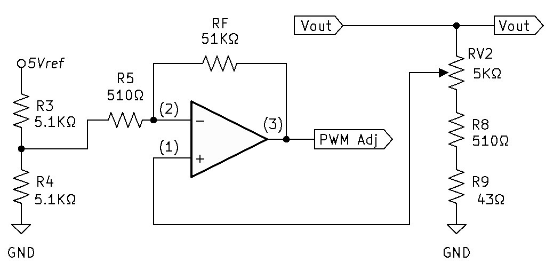


Figure 12b.14: TL494 Variable Voltage

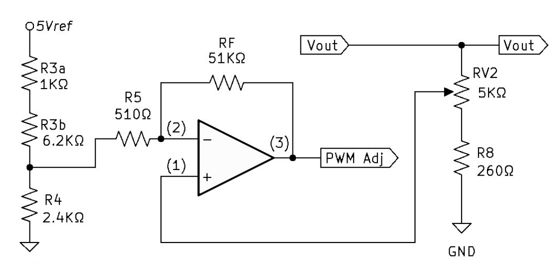


Figure 12b.15: TL494 Variable Voltage

12b.5.4 TL494 Circuit Schematics

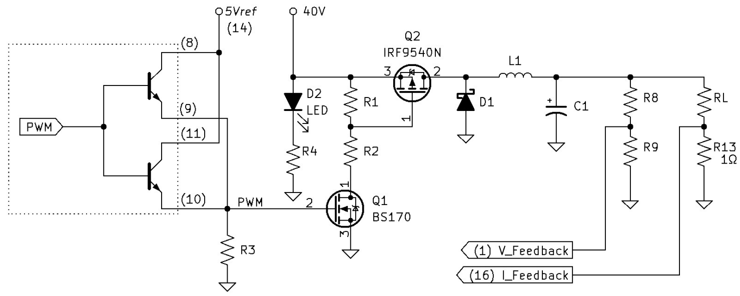


Figure 12b.16: TL494 Output PWM to BS170

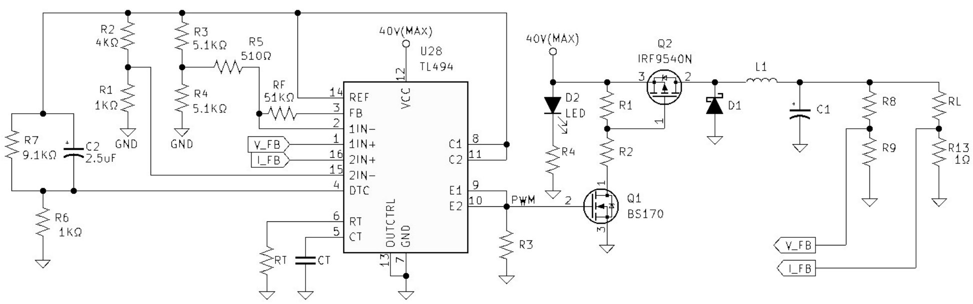


Figure 12b.17: TL494 Schematic Fixed Voltage and Current Regulation

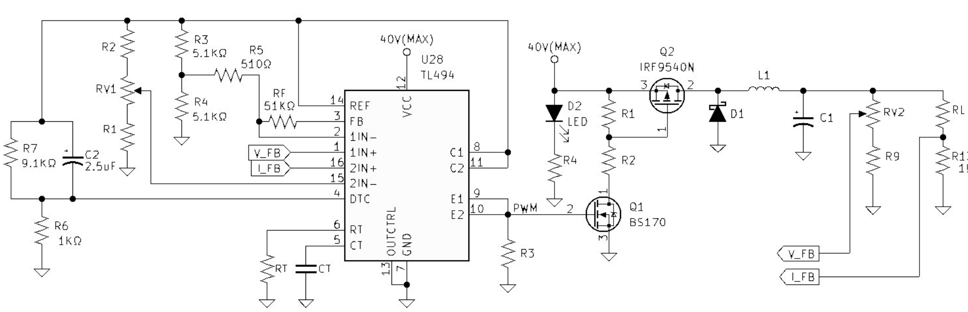


Figure 12b.18: TL494 Schematic Variable Voltage and Current Regulation

12b.6 MC3406ACN Integrated SMPS

12b.6.1 Datasheet

Download datasheet here: MC3406ACN Datasheet

- Step-up, Step-down, and Inverting applications
- Quiescent current 2.5mA (typ.)
- 3 V to 40 V
- Frequency operation 100 kHz
- Active current limiting

12b.6.2 Buck

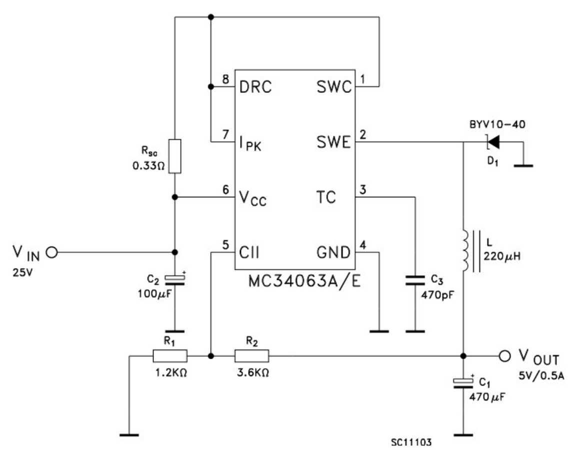


Figure 12b.19: MC3406ACN Buck

Vout Calculated:

- $V_{out} = \frac{V_{th}}{R_1}(R1 + R2) = \frac{1.25V}{1200\Omega}(1200\Omega + 3600\Omega)$
- $V_{out} = 5V$

Iout Current Limit Calculated:

- $Iout_{CurrentLimit} = \frac{V_{IPK(sense)}}{R_{SC}} = \frac{0.3V}{0.33\Omega}$
- $Iout_{CurrentLimit} = 909mA$

12b.6.3 Boost

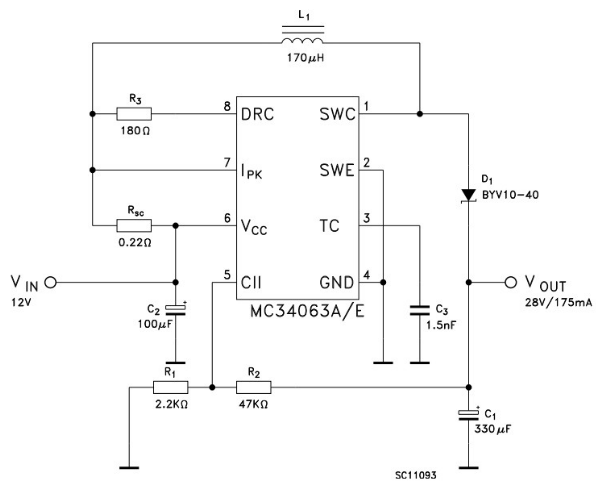


Figure 12b.20: MC3406ACN Boost

Vout Calculated:

- $V_{out} = \frac{V_{th}}{R_1}(R1 + R2) = \frac{1.25V}{2200\Omega}(2200\Omega + 47000\Omega)$
- $V_{out} = 27.95V$

Iout Current Limit Calculated:

- $Iout_{CurrentLimit} = \frac{V_{IPK(sense)}}{R_{SC}} = \frac{0.3V}{0.22\Omega}$
- $Iout_{CurrentLimit} = 1.364A$

12b.6.4 Buck-Boost Inverter

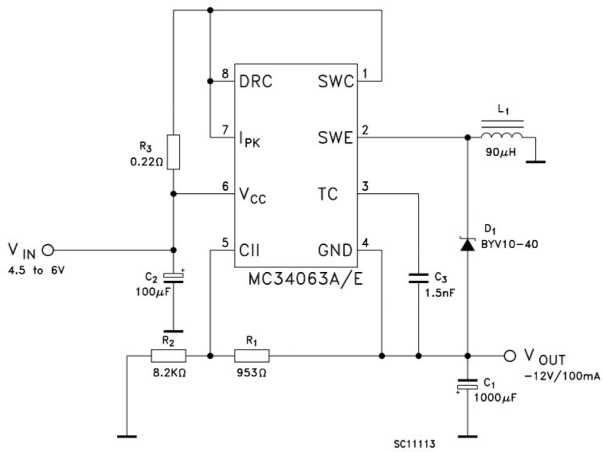


Figure 12b.21: MC3406ACN Buck-Boost Inverter

Vout Calculated:

- $V_{out} = \frac{V_{th}}{R_1}(R1 + R2) = \frac{1.25V}{953\Omega}(953\Omega + 8200\Omega)$
- $V_{out} = -12V$

Iout Current Limit Calculated:

- $Iout_{CurrentLimit} = \frac{V_{IPK(sense)}}{R_{SC}} = \frac{0.3V}{0.22\Omega}$
- $Iout_{CurrentLimit} = 1.364A$

12b.7 LM2576T Integrated SMPS

12b.7.1 Datasheet

Download datasheet here: LM2576T Datasheet

- Step-down BUCK SMPS
- 3.3-V, 5-V, 12-V, 15-V, and adjustable output versions
- Specified 3-A output current
- Wide input voltage range: 40 V Up to 60 V for HV version
- Requires only four external components

12b.7.2 Fixed 5V

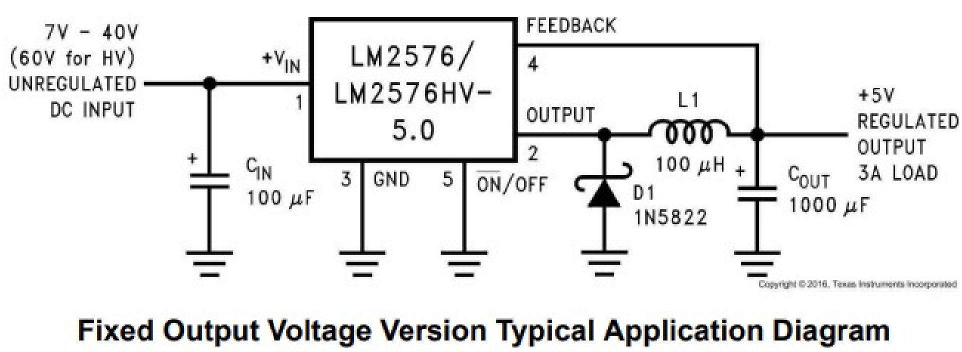


Figure 12b.22: LM2576T-5 Fixed 5V

12b.7.3 Adjustable

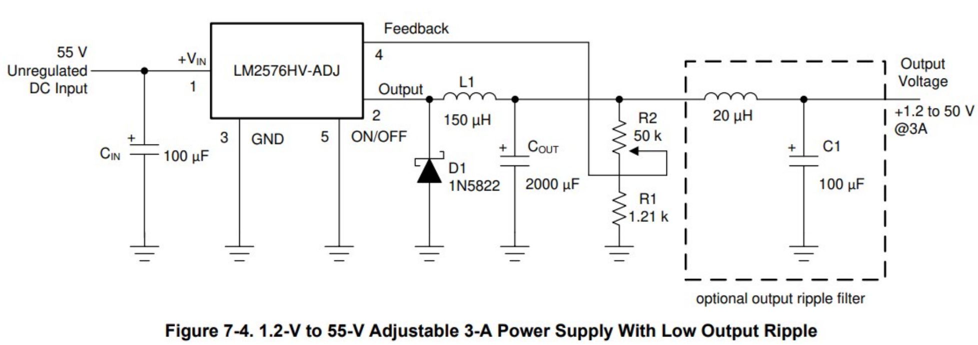


Figure 12b.23: LM2576T-Adj

12b.7.4 Schottky Diode Guide

Table 8-3. Diode Selection Guide

V_R	SCHOTTKY		FAST RECOVERY	
	3 A	4 A to 6 A	3 A	4 A to 6 A
20 V	1N5820 MBR320P SR302	1N5823	The following diodes are all rated to 100-V 31DF1 HER302	The following diodes are all rated to 100-V 50WF10 MUR410 HER602
30 V	1N5821 MBR330 31DQ03 SR303	50WQ03 1N5824		
40 V	1N5822 MBR340 31DQ04 SR304	MBR340 50WQ04 1N5825		
50 V	MBR350 31DQ05 SR305	50WQ05		
60 V	MBR360 DQ06 SR306	50WR06 50SQ060		

Figure 12b.24: Schottky Diode Guide

12b.7.5 Inductor Guide

Table 8-4. Inductor Selection by Manufacturer's Part Number

INDUCTOR CODE	INDUCTOR VALUE	SCHOTT[1]	PULSE ENG.[2]	RENCO[3]
L47	47 µH	671 26980	PE-53112	RL2442
L68	68 µH	671 26990	PE-92114	RL2443
L100	100 µH	671 27000	PE-92108	RL2444
L150	150 µH	671 27010	PE-53113	RL1954
L220	220 µH	671 27020	PE-52626	RL1953
L330	330 µH	671 27030	PE-52627	RL1952
L470	470 µH	671 27040	PE-53114	RL1951
L680	680 µH	671 27050	PE-52629	RL1950
H150	150 µH	671 27060	PE-53115	RL2445
H220	220 µH	671 27070	PE-53116	RL2446
H330	330 µH	671 27080	PE-53117	RL2447
H470	470 µH	671 27090	PE-53118	RL1961
H680	680 µH	671 27100	PE-53119	RL1960
H1000	1000 µH	671 27110	PE-53120	RL1959
H1500	1500 µH	671 27120	PE-53121	RL1958
H2200	2200 µH	671 27130	PE-53122	RL2448

Figure 12b.25: Inductor Guide

12b.8 LM2677 Integrated SMPS

12b.8.1 Datasheet

Download datasheet here: LM2677 Datasheet

- Step-down BUCK SMPS
- 3.3-V, 5-V, 12-V, and adjustable output versions
- Specified 5-A output current
- Wide input voltage range: 8 V to 40 V
- Efficiency up to 92%

12b.8.2 Fixed 5V

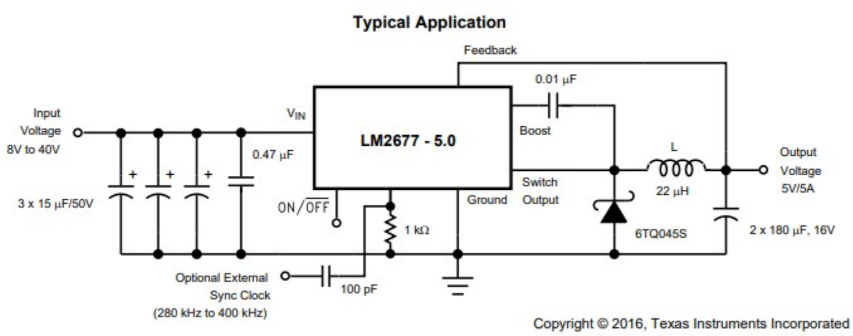


Figure 12b.26: LM2677-5 Fixed 5V

12b.8.3 Adjustable

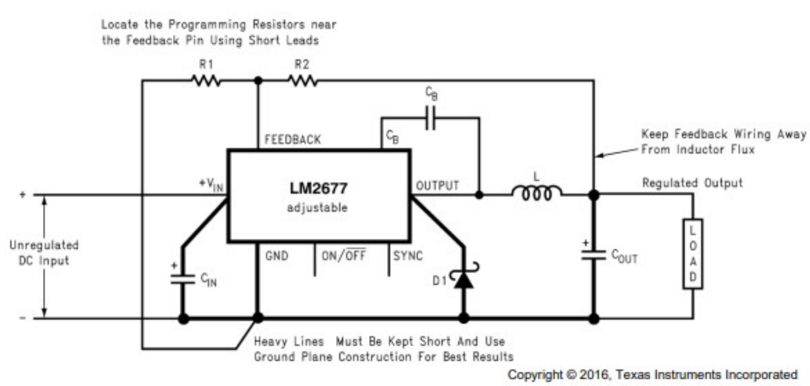


Figure 12b.27: LM2677-Adj

12b.8.4 Schottky Diode Guide

Table 7. Schottky Diode Selection Table

REVERSE VOLTAGE (V)	SURFACE MOUNT		THROUGH HOLE	
	3 A	5 A OR MORE	3 A	5 A OR MORE
20	SK32	—	1N5820	—
		—	SR302	—
30	SK33	MBRD835L	1N5821	—
	30WQ03F		31DQ03	—
40	SK34	MBRB1545CT	1N5822	—
	30BQ040	6TQ045S	MBR340	MBR745
	30WQ04F	—	31DQ04	80SQ045
	MBRS340	—	SR403	6TQ045
	MBRD340	—	—	—
50 or more	SK35	—	MBR350	—
	30WQ05F	—	31DQ05	—
	—	—	SR305	—

Figure 12b.28: Schottky Diode Guide

12b.8.5 Inductor Guide

Table 4. Inductor Manufacturer Part Numbers[0] (continued)

INDUCTOR REF. #	INDUCTANCE (µH)	CURRENT (A)	RENCO		PULSE ENGINEERING		COILCRAFT
			THROUGH HOLE	SURFACE MOUNT	THROUGH HOLE	SURFACE MOUNT	SURFACE MOUNT
L34	15	3.65	RL-1283-15-43	—	PE-53934	PE-53934S	DO5022P-153
L38	68	2.97	RL-5472-2	—	PE-54038	PE-54038S	—
L39	47	3.57	RL-5472-3	—	PE-54039	PE-54039S	—
L40	33	4.26	RL-1283-33-43	—	PE-54040	PE-54040S	—
L41	22	5.22	RL-1283-22-43	—	PE-54041	P0841	—
L44	68	3.45	RL-5473-3	—	PE-54044	—	—
L45	10	4.47	RL-1283-10-43	—	—	P0845	DO5022P-103HC
L46	15	5.60	RL-1283-15-43	—	—	P0846	DO5022P-153HC
L47	10	5.66	RL-1283-10-43	—	—	P0847	DO5022P-103HC
L48	47	5.61	RL-1282-47-43	—	—	P0848	—
L49	33	5.61	RL-1282-33-43	—	—	P0849	—

Figure 12b.29: Inductor Guide

Week 13

Relays, Solenoids, Motors and Servos

13.1 What is an Actuator Video

- What is an Actuator?

13.2 My Military Notes:

13.2.1 Relay and Solenoid Operation

RELAY/SOLENOID OPERATION

OBJECTIVES

Identify relay operating principles.

Identify solenoid operating principles.

INTRODUCTION

In this lesson, the basic operation of the mechanical relay and solenoids will be covered. The mechanical relay works by applying a voltage to a coil, therefore, it is sometimes known as an electromechanical relay. Since a coil can become a magnet, another term commonly used is electromagnetic relay. There are many different types and purposes for relays; however, the basic principles of operation and structural features are similar. Relays are used in electronic equipment for many different reasons, but basically, they act as an electronically controlled switch used to change voltage/current levels within the equipment. A solenoid is a control device that uses electromagnetism to convert electrical energy into mechanical motion. The movement of the solenoid may be used to close a set of electrical contacts, cause the movement of a mechanical device or both.

INFORMATION

RELAYS

Relay Operating Principles

Before looking at the schematics, it is necessary to explain a few terms used when discussing relays. "Pole" refers to the movable contact of a relay. If a relay is a single-pole, it means it has only one movable contact. "Throw" refers to the number of switch options available. Single throw means there is only one switch to be closed. "Break" means the contacts have been opened by either mechanical or electrical action. "Make" means the contacts are touching each other and a complete circuit has been formed.

Symbols. Relays are represented by schematic symbols as shown in Figure 1. Figure 1-A shows a single-pole, single-throw (SPST) relay with the contacts normally open (NO). Figure 1-B shows a SPST relay with the contacts normally closed (NC). When current flows thru the coil, the contacts open. Figure 1-C shows a single-pole, double-throw (SPDT) relay with one contact NC and one NO. This type of relay is used to transfer current from one circuit to another. A single relay may have multiple poles with any combination of opened or closed contacts (refer to Figure 1-D).

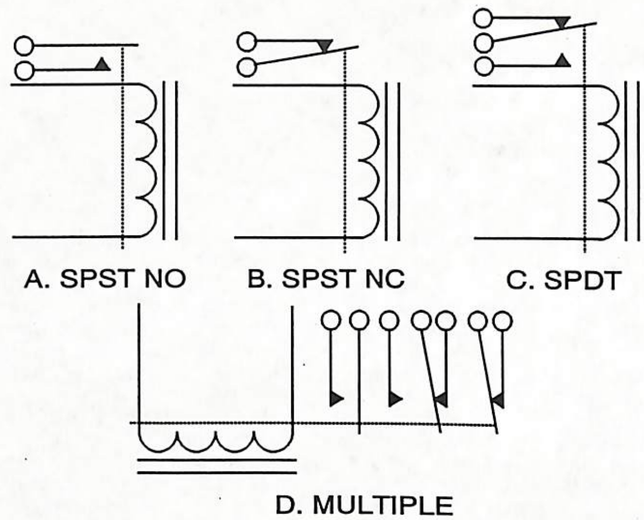


Figure 1. Schematic Symbols

The relays shown in Figure 1 are in the de-energized (unmagnetized) condition; no current is flowing through the coil. In all schematic diagrams you will see, the relays are always shown in the de-energized condition unless otherwise indicated.

Relays are electromagnetic devices that are used for controlling or switching electrical circuits from some remote location. The circuit that energizes the relay is called the actuating control, or primary circuit. The circuits that are switched (open or closed) are called the controlled or secondary circuits.

Construction. A relay consists of five main parts. These parts are illustrated in Figure 2. An explanation of each part follows.

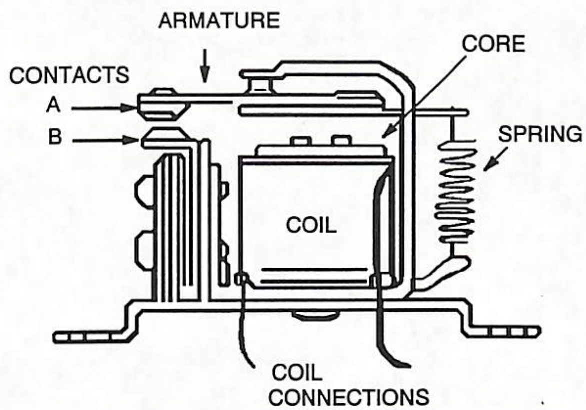


Figure 2. Basic Relay

Armature. This is the moving part of the relay that operates the contacts. (See A in Figure 2). The armature of most relays is constructed on the principle of the lever. It is made stationary on one end and pivots through an arc on its mounting.

Contacts. These are used to open or close the relay secondary circuits. Contacts may be of the movable or stationary types. The stationary contact (B in Figure 2) is mounted on the base and insulated from the rest of the relay. The contacts can be made of gold, silver, silver alloys, or platinum alloys, and often are plated with cadium or nickel. Relay contacts are rated by maximum current handling capabilities and voltage ratings. These ratings depend upon the types of contact material, size, and spacing of the open contacts. It is very important not to exceed these ratings, as the contacts will arc and become worn or pitted. In this condition, they can not function properly.

Coil. When power is applied to the coil connections, the current through the coil causes a strong magnetic field around the coil. This causes the armature to be attracted to the core. Coils are rated according to voltage and power handling capabilities.

Core. When the magnetic field builds up around the core, it attracts the metal material of the armature and pulls it toward the core.

Spring. This provides the constant restraining force that keeps the armature in the normal or de-energized position (Figure 2). (De-energized describes a relay without voltage applied to the coil as opposed to an energized relay with voltage applied). When the relay is energized the spring opposes the motion of the armature as it moves into the closed position. When the relay becomes de-energized, the spring returns the armature back to the open position.

Figure 4 shows a holding relay circuit. Switches S1 and S2 are spring loaded switches which return to their original positions as soon as they are released. S1 is normally closed and S2 is normally open. When the relay is de-energized the contacts do not provide any current paths, thus the lamp will not light. Momentarily closing switch S2 will permit current to flow through the coil and the lamp. The armature will be attracted toward the electromagnet, thus closing relay contact A and providing a current path. This provides an additional path for current through the coil and the lamp. Releasing or opening switch S2 will not stop current through the coil or lamp. The relay contact will keep or hold the circuit closed. In order to stop current flow, switch S1 must be momentarily opened.

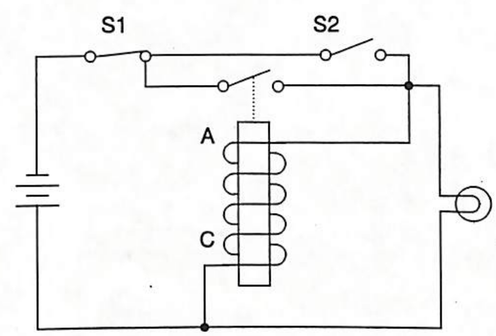


Figure 4. Holding Relay

Opening S1 will cause the relay to de-energize and open contact A. Releasing or (closing) S1 will not re-establish current. S2 must once again be closed to turn on the lamp and energize the relay. S2 is sometimes marked ON and S1 marked OFF.

Figure 5 shows a starting relay circuit. A common use of this relay is in an automobile starter circuit. The ignition switch is a momentary switch that is normally open, thus the relay will only be energized as long as this switch is held closed. In the relay's de-energized condition, contact B (an iron bar) is open so the current path to the starter motor is open. When the ignition switch is held closed, current flows through the relay coil. This causes an electromagnetic force which pulls the iron bar B toward the coil and closes the starter motor circuit. This provides a path for a large current to flow in the starter motor circuit. Note that the conductors in this circuit are short, heavy cables. A small current in the relay coil circuit controls a very large current in the starter circuit. The relay will remain energized until the ignition switch is released.

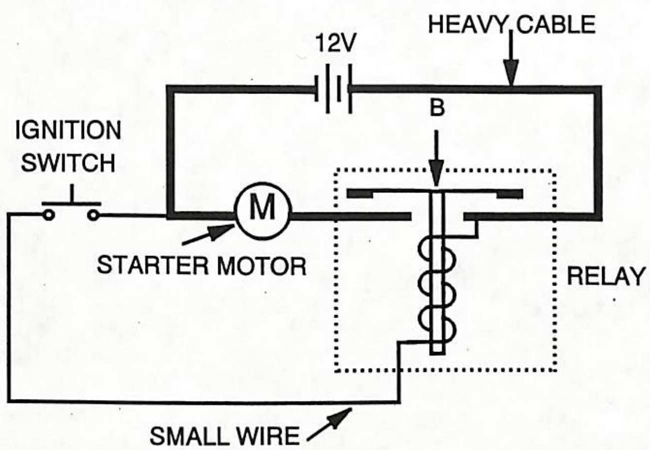


Figure 5. Starting Relay

SOLENOIDS

Solenoid Operating Principles

A solenoid, like a relay, is a control device that uses electromagnetism to convert electrical energy into mechanical motion. While relays are used strictly to close a set of electrical contacts, solenoids may be used to close a set of electrical contacts, cause the movement of a mechanical device, or both.

Purpose. While a solenoid is a control device, some other control device, such as a switch or a relay, controls the solenoid. One advantage of a solenoid is that it can control mechanical movement at a considerable distance from the control device. The only link necessary between the control device and the solenoid is the electrical wiring for the coil current. Solenoids can be used to actuate large contacts for the control of high current. Therefore, a solenoid, like a relay, can provide a means of controlling high current with a low current switch.

Construction. Figure 6 is a cutaway view of a solenoid showing the solenoid action. A solenoid is a electromagnet formed by a conductor wound in a series of loops in the shape of a spiral. Inserted within this coil is a soft-iron core and a movable plunger. The soft-iron core is pinned or held in a stationary position. The movable plunger (also soft-iron) is held away from the core by a spring when the solenoid is de-energized.

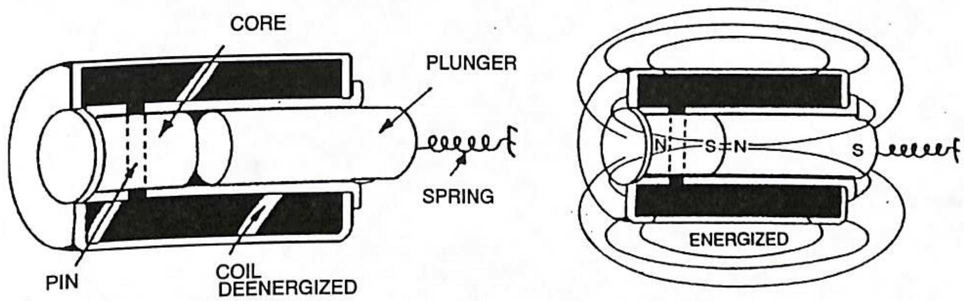


Figure 6. Solenoid Operation

Operation. When current flows through the conductor, it produces a magnetic field. The magnetic flux produced by the coil induces north and south poles in both the core and plunger. The plunger produces mechanical motion as it is magnetically moved into the core. An example of this would be the starter solenoid on an automobile.

As shown in Figure 6, the de-energized position of the plunger is partially out of the coil due to the action of the spring. When the coil is de-energized, the plunger returns to this normal position. The effective strength of the magnetic field on the plunger varies according to the distance between the plunger and the core. The strength of a magnetic field is inversely proportional to the distance from the core. As distance increases, the magnetic field strength decreases.

SUMMARY

Relays and solenoids depend on the effects of electromagnetism to operate. Relays are used throughout electronics as electronically controlled switches to control associated circuitry. Solenoids are used to convert electrical energy into mechanical motion to close electrical contacts, cause the movement of a mechanical device, or both.

Refer to Safety Precautions Checklist before performing trainer project.

13.2.2 Relay and Solenoid Fault Isolation

RELAY/SOLENOID FAULT ISOLATION

OBJECTIVE

Given test equipment and a trainer, isolate a faulty relay/solenoid in a circuit.

INTRODUCTION

Your job as an electronics technician may at some time involve finding and replacing relays/solenoids that are malfunctioning. In this lesson you will learn what can go wrong with relays/solenoids and how to isolate a faulty relay in a circuit. You will also troubleshoot a circuit to isolate a faulty relay.

INFORMATION

FAULT ISOLATION TECHNIQUES

The main problems found with relays involve the coil and the contacts. If the coil burns open there will be no magnetic field to move the armature, thus the relay will be rendered inoperative. Every time a pair of contacts moves apart a spark occurs. The same thing happens every time two contacts come together. This sparking, called "arcing", can eventually cause the contacts to burn and become pitted, resulting in a poor connection that interrupts the path for current in the controlled circuit.

Visual Inspection

The type of relay used in a circuit, sealed or unsealed, will determine how useful a visual inspection will be when fault isolating. A visual inspection of a sealed relay is limited mainly to checking for obvious physical damage or loose connections. Unsealed relays have a removable cover which allows for a closer and more complete visual inspection.

Armature. The moving armature in a relay can cause malfunctions due to the fact that any moving part may eventually wear out. Armatures can break or stick in the energized or de-energized position.

Spring. The spring that supplies the restoring force can break, lose its tension, or be out of adjustment. These problems can cause the relay to "chatter" or to stay in its energized position even if the current flow through the coil is turned off.

Contacts. Contacts can become pitted or deposits may build up on the contacts causing a poor electrical connection. If the contacts are pitted then the relay should be replaced unless new contacts can be fitted. If the inspection reveals an oxide coating or carbon deposit, use a contact "burnishing tool" to wipe the contacts clean. Never file the contacts.

Voltage Checks

If a relay is believed to be the cause of a circuit not working properly, then the relay can normally be checked with a multimeter. If the power is on, voltage checks can be made. If the power is removed or the relay is out of the circuit, then resistance checks can be made.

In many pieces of military equipment, the relay coil is connected to a 28 volt DC source, as shown in Figure 1. When the switch is closed, the path for current is complete, and current should flow through the coil. To make sure that the switch and the connecting wires are good, a voltmeter can be connected to the coil connections (3 and 7). If a reading of 28 volts is indicated, then the primary power is getting to the relay.

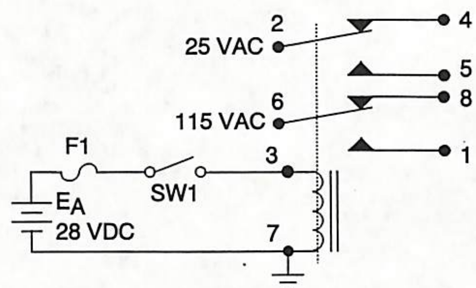


Figure 1. Relay Contact Current Paths

Even if the coil is getting its voltage, the relay may still malfunction. In Figure 1, 115 volts AC is present at the lower wiper (contact 6). If the relay energizes properly, then the 115 volts should be present at contact 1. If a voltmeter indicates that the 115 volts is not present at contact 1, then either the armature is not moving, or there is a poor connection between the switching contacts. Since the relay is supposed to be energized, the voltage at the de-energized contact 8 should read 0 volts on a voltmeter. A 115 volt reading would indicate that the relay is not energizing correctly.

The upper wiper (contact 2) has 25 volts AC applied to it. If the relay is supposed to be energized, then the same 25 volts AC should also be present at contact 5. A voltmeter connected to contact 5 should indicate 25 volts AC. There should be 0 volts at contact 4. If the voltmeter indicates 0 volts at contact 5 and 25 volts at contact 4, then the relay is not energizing.

If the relay were de-energized (SW1 open), then a voltmeter would be expected to indicate 115 volts AC at contacts 6 and 8, but 0 volts at contact 1. 25 volts AC would be expected at contacts 2 and 4, with a 0 volt reading at contact 5. Improper readings would indicate a faulty relay.

Resistance Checks

Resistance checks with an ohmmeter are usually performed with the power removed from the entire circuit, and with the relay isolated from the rest of the circuit. A resistance check of the coil should read in the 100 to 400 ohm range on most relays.

A resistance check of the de-energized contacts (2 and 4 or 6 and 8 in Figure 1) should indicate continuity (0 ohms) with the relay off and disconnected. A resistance check of the energized contacts (2 and 5 or 6 and 1 in Figure 1) should indicate discontinuity (infinite resistance). Resistance checks of the switching contacts can safely be done with power applied to the coil in some situations, but you will not be required to perform any during this lesson.

If a relay is discovered to be malfunctioning, it is usually replaced because most relays are sealed units and cannot be repaired. Replacement of a relay must be done carefully to avoid incorrect or shorted connections which could cause damage to the equipment you are trying to repair.

Solenoids

Solenoids are very similar to relays and thus are subject to many of the same types of malfunctions. The plunger in a solenoid, because it is a moving part, can cause malfunctions similar to those caused by the armature in a relay. Plungers can break or stick in the energized or de-energized position. The spring that supplies the restoring force can break or lose its tension causing the plunger to stay in its energized position even if the current flow through the coil is turned off.

SUMMARY

Fault isolating relays/solenoids should be done by conducting a visual inspection and then voltage and resistance checks.

Refer to Safety Precautions Checklist before performing trainer project.

13.2.3 DC Motor Operation

DC MOTOR OPERATION

OBJECTIVE

Identify DC motor operating principles.

INTRODUCTION

DC motors are used to convert electrical energy into mechanical energy. The electrical energy develops a magnetic field which interacts and produces a mechanical force. Motors come in many types and sizes to perform different amounts of work.

INFORMATION

Construction

The major parts in a practical DC motor are the armature assembly, the field assembly, the brush assembly and the end of frame. See Figure 1.

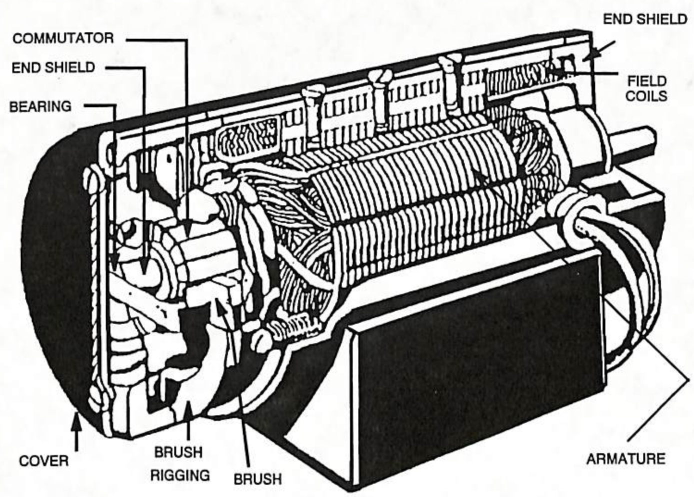


Figure 1. Parts of a DC Motor

Armature assembly. The armature assembly contains: a laminated soft iron core, coils, and a commutator. The core, coils, and commutator are all mounted on a rotating steel shaft. (Figures 1 and 2)

Laminated soft iron core. Laminations (stacks of soft iron) insulated in slots which are insulated from each other form the armature core revolving in the magnetic field.

Armature windings. Insulated copper wire windings which are inserted in slots which are insulated with fiber paper (fish paper) to protect the windings.

Commutator. Consists of a large number of copper segments insulated from each other and the armature shaft by pieces of mica insulated wedge rings that hold the segments in place.

The laminated soft iron core covers and surrounds the majority of the steel shaft. The armature windings are wound around the iron core and inserted into slots to protect the windings. Wedges, or steel bands, hold the windings in place to prevent them from flying out of the slots when the armature assembly is rotating at high speeds. The commutator segments are connected to the ends of the armature windings.

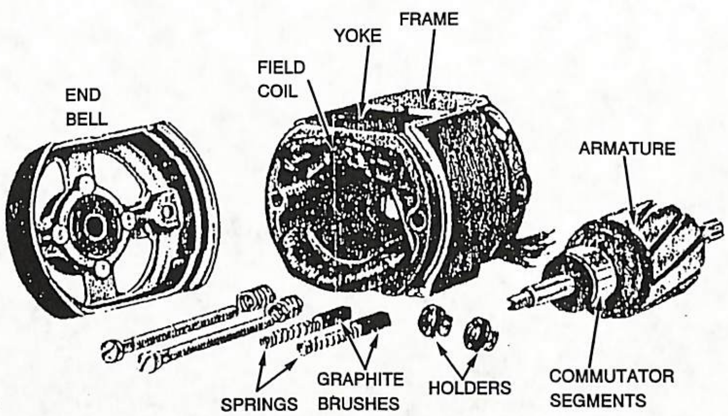


Figure 2.

Field assembly. The field assembly may consist of: the field frame, pole pieces, and the field coils as seen in Figure 1. (Note: The field assembly may include only a simple permanent magnet as will be discussed in the Operation of DC motors. Either field assembly is acceptable and operable.)

Field Frame. Located along the inner wall of the motor housing and contains pole pieces and field coils.

Pole pieces. Laminated and soft steel around which the field coils are wound.

Field coils. Consists of several turns of insulated wire which are wrapped around the pole pieces and housed within the field frame.

A magnetic field develops when the field coils have current flowing through them. This magnetic field from the field coils interacts with the magnetic field from the armature coils to cause the armature assembly to rotate.

Brush assembly. The brush assembly consists of: the brushes, springs, and their holders. Figure 1 shows where the brushes are located. Figure 2 indicates the different parts of the brush assembly.

Brushes. Small blocks of graphitic carbon (allows long service life and causes minimum wear to the commutator).

Springs. Holds the brushes firmly against the commutator.

Holders. Permits some movement in the brushes so they can follow any irregularities in the surface of the commutator and continue to maintain good contact.

Current is conducted from the power source through the brushes to the commutator segments which allows conduction to continue on to the armature windings. Therefore, the brushes may be considered to be the electrical connection from the power source to the internal circuitry of the DC motor.

End frame. The end frame, or endbell, is the part of the motor opposite the commutator. It supports the bearing and protects the electrical and rotating parts inside the motor. Sometimes the end frame is designed to be part of the unit driven by the motor. Figure 2 shows the end frame.

Operation

Laws of magnetism. DC Motors operate on the principle: a force is exerted between stationary magnetic fields (brought about by magnetic field from field coils or permanent magnet) and movable magnetic fields (magnetic field from armature coils). The amount and direction of this force determines motor speed and direction of rotation. In order to determine the amount of force and its direction, the strength and polarity of both magnetic fields must be known.

Distortion of a magnetic field by a current-carrying conductor (current flows out of paper)

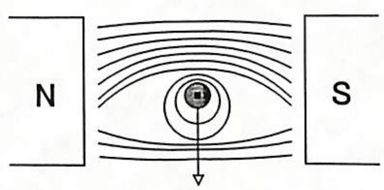


Figure 3.

When a current-carrying wire is placed in a magnetic field, a force acting on the wire is set up at right angles to the magnetic field. The lines of force in the field from the permanent magnet are exerted from the north to south pole. When no current flows through the wire, no force is exerted upon it. However, when current flows through the wire, a magnetic field is set up around it. Using the left hand rule, it is evident that the magnetic field develops in a clockwise direction as shown in Figure 3. The field around the wire interacts with the magnetic field between the north and south poles permanent magnet causing the wire to feel a downward force being applied to it and thereby moving downward.

Distortion of a magnetic field by a current-carrying conductor (current flows into paper)

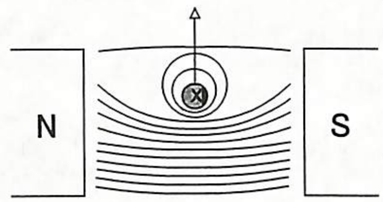


Figure 4.

Another current-carrying wire may have current flowing in the opposite direction, into the paper. Using the left hand rule indicates that the magnetic field develops around the wire in a counter-clockwise fashion as shown in Figure 4. The field around the wire interacts with the magnetic field between the permanent magnet causing the current-carrying wire to feel an upward force being exerted onto it and thereby causing it to move upward.

Cross section of a single loop (armature coil) carrying current in a magnetic field

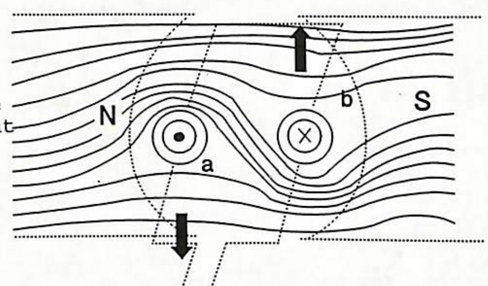


Figure 5.

The single wire can be replaced by a single loop (a cross section of a single loop-armature coil- of a conductor) carrying current and developing magnetic fields in opposite directions based upon the direction in which the current is flowing through the loop as seen in Figure 5. The current is flowing out of the paper at **a** and into the paper at **b** as shown. This causes the magnetic field to be strengthened above **a** and below **b**. Thus, there is a force pushing down on a and up on b. If the loop is free to rotate, it will do so in a counterclockwise direction. This force, that produces rotation in either a clockwise or counterclockwise direction, is called <u>torque</u>. The single loop represents a simple armature. In a practical motor, however, many loops of wire are wound around an iron core to form the armature.

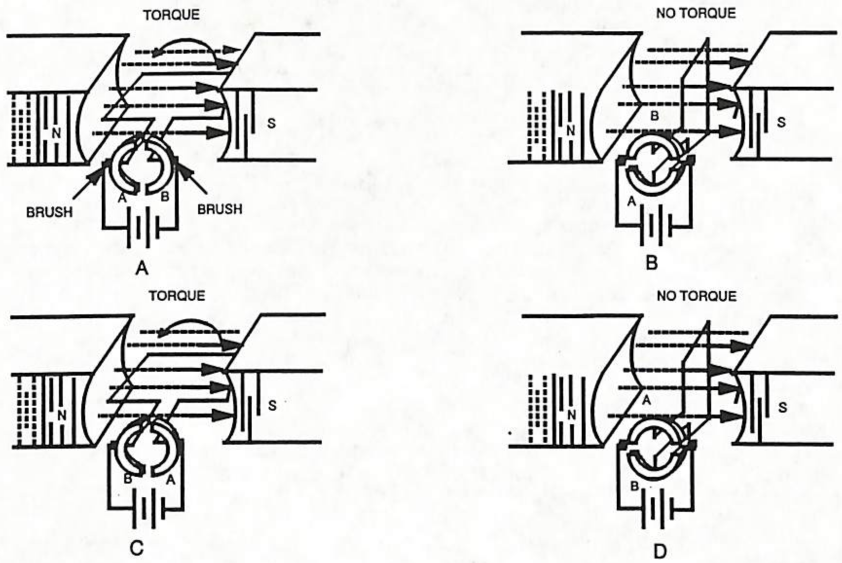


Figure 6.

Torque. As discussed earlier, a coil of wire (single loop) through which current flows will rotate when placed in a magnetic field. Figure 6 represents a coil mounted in a magnetic field in which it can rotate. If the connecting wires from the battery were permanently fastened to the terminals of the coil, when current flowed, the coil would rotate only until it aligned itself with the magnetic field, then stop. Torque at that point would be zero. However, a motor must keep rotating, so it becomes necessary to change the direction of current flow through the coil. As the plane of the coil becomes parallel to the lines of force, a change of commutator segments takes place and the direction of current flow is in the opposite direction. This allows the coil to continue rotating.

Rotating Coil. In Figure 6, the terminals of the coil are labeled A and B. As the coils rotate, they slide against the fixed terminals (brushes). The direction of current flow in the side of the coil next to the north-seeking pole is outward; the force acting on that side of the coil turns it downward. The commutator changes the current from one wire to another; the segments (A and B in Figure 6A) are connected to the ends of the revolving loop.

With the coil positioned as shown in Figure 6A, current flows from the negative terminal of the battery to the negative brush to segment B of the commutator through the loop of segment A, to the positive brush and back to the positive terminal of the battery. The coil rotates counterclockwise. Torque is maximum, as the greater number of lines of force are being cut by the coil.

When the coil has rotated 90° to the position in Figure 6B, segments A and B of the commutator no longer make contact with the battery circuit, and no current flows through the coil. Torque has minimum value, as a minimum number of lines of force are being cut. The momentum of the coil carries it beyond this point until the segments again make contact with the brushes and current again flows through the coil. This time it flows into segment A and out of segment B (Figure 6C). The positions of segments A and B also have been reversed so the effect of the current is the same as before. The torque acts in the same direction, and the coil continues to rotate in its counterclockwise direction. When the coil passes through the position in Figure 6C, torque again reaches maximum. Continued rotation carries the coil again to a position of minimum torque (Figure 6D). At this position, the brushes no longer carry current; once more, momentum rotates the coil to the point where current enters through segment B and leaves through A. Further rotation returns the coil to the starting point, and one revolution is completed.

The torque in a motor containing only a single coil is neither continuous nor very effective as there are two positions where there is actually no torque at all. To overcome this, a practical DC motor contains a large number of coils wound on the armature. These coils are spaced so that for any position of the armature there will be coils near the poles of the magnet. This makes the torque both steady and strong. The commutator contains a large number of segments instead of just two.

In a practical motor, the armature assembly is placed between the poles of an electromagnet (Field coils), which provides a strong magnetic field. The core, usually mild or annealed steel, can be magnetized strongly through induction. The magnetizing current is from the same source that supplies the current to the armature.

Types of Motors and Characteristics

Motors are normally classified according to the voltage or current used (AC or DC), and by the method of motor excitation. An example of uses for DC motors would be battery operated devices (drills, shavers, toys, or something as large as a golf cart).

Series. The series motor is so named because the armature and field windings are connected in series, as shown in Figure 7. This method of connection makes it necessary for the field to be heavy enough to carry the armature current. Consequently, the wire in a series field is of a large size, and the winding contains relatively few turns. The one serious disadvantage of the series motor is that the speed of the motor varies inversely with the load. Should the load be removed, the increase in motor speed could result in destruction of the motor. Series motors are not operated without a load. Therefore, this type of motor is not adaptable to a constant-speed application, although it is ideal for use where the speed of the motor is continuously under the control of an operator, and where a high starting torque is required. In most small series motors (electric drills for example) the gears load the motor enough to prevent motor runaway. The torque of a series motor varies as the square of the current varies. (The torque will increase as the square of the current increases up to the point were the core material reaches saturation.) Because of its speed characteristics, the series motor is not suitable for any load where the required torque might drop below 15 percent of full-load torque.

Direction of rotation for a series motor may be changed by:

1. Reversing current flow in the field winding.

2. Reversing current flow in the armature.

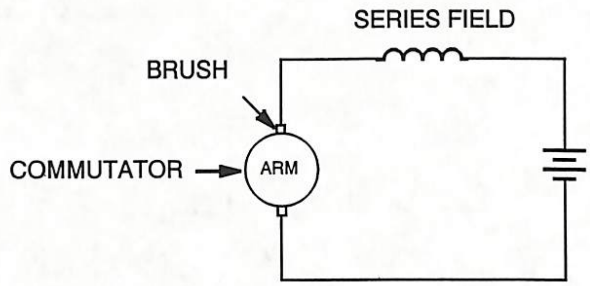


Figure 7. Series Motors.

Shunt. In the shunt-motor, the field winding is connected in parallel with the armature, as shown in Figure 8.

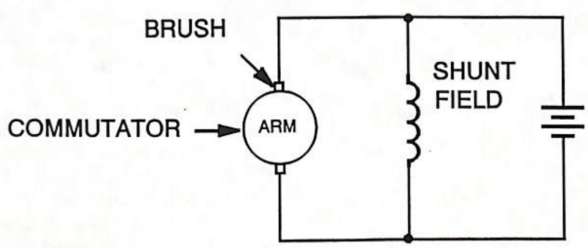


Figure 8. Shunt Motor.

There are two paths for current in the shunt motor, one through the armature and one through the field. The field coils are wound with relatively small wire, and have a large number of turns. This type of winding is necessary to maintain a high flux value and to prevent coil overheating. Because the field is connected across the line, the density of the magnetic field remains constant. Therefore, the torque of a shunt motor must vary with the current in the armature; that is, if the armature current is doubled, the torque is also doubled. Since the field strength is constant, the motor speed will be relatively constant (+ 1 percent) from no load to full load. The shunt-wound motor is the type used for constant-speed applications, but, because of the fixed field current, it has a lower starting torque than the series motor. The lower starting torque indicates that if a series and shunt motor were equivalent in size of horsepower, the series motor produces the greater starting torque.

Direction of rotation for a shunt motor is changed in the same manner as in the series motor. Reversing current flow in either the armature or in the field coil changes direction of rotation.

Compound. When the series and shunt-type windings are combined in one motor, the resulting machine has both high starting torque and constant speed. See Figure 9.

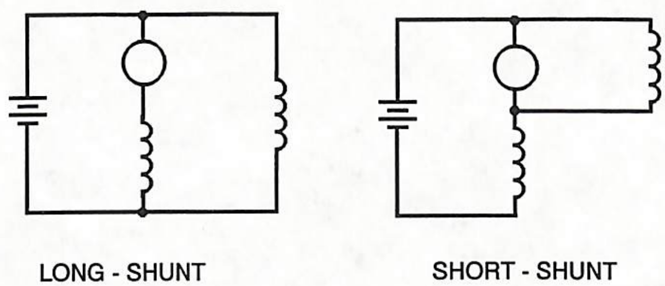


Figure 9. Compound Motor.

In a shunt motor, the series field provides more torque as the load increases. When the load increases, the motor slows down; this results in an increase in armature current which boosts the strength of the series field winding. With more total interacting flux, torque increases, and the motor regains its speed.

In a series motor, the shunt field provides constant flux, and thus constant speed. The tendency of the series motor to run away with no load now has a counterforce of a constant flux field. Thus, motor speed limits itself to a reasonable value, as in the case of the regular shunt motor.

In some applications, the series winding is in the circuit for starting only, and is switched out when the motor is up to normal speed. The compound-type motor does not have a flat constant speed under load, but its variation is not great (between 15 and 20 percent). This variation is dependent upon the number of ampere-turns in the series winding; the greater the number of ampere-turns, the greater the starting torque, and the greater the speed variation (unless the series winding is switched out when the motor reaches normal speed). As the ampere-turns of the series winding are decreased, the speed variation under changing load is decreased. With a weakened shunt field, the series-field flux constitutes the greater portion of the total flux; hence, changes in load may produce unstable speed.

Ratings

DC motors can be rated by voltage, RPM, or speed control. These are defined below. Be sure to check the data on the ratings plate to ensure the motor is suitable for the intended application.

Voltage. For a circuit or device, the recommended maximum voltage which may be applied, or the recommended working voltage, as specified.

Horsepower (hp). Abbreviated hp. A unit of power, or the capacity of a mechanism to do work.

RPM. The rate of rotation of the shaft of the motor. This is usually referred to as "revolutions per minute".

Speed Control. DC motor speed is normally controlled by current through the armature, which is dependent upon the armature resistance, field strength, or application.

SUMMARY

Major assemblies of the DC motor were identified. The operation of the DC motor was covered next. This was followed by the types and characteristics of the series, shunt, and compound motors. The last area to be covered identified the different methods used in determining the motor rating.

13.2.4 AC Motor Operation

AC MOTOR OPERATION

OBJECTIVE

Identify AC motor operating principles.

INTRODUCTION

Motors are electromechanical devices that perform important functions in electrical applications. Motors convert electrical energy into mechanical energy. In this lesson we will be concerned with AC, or alternating current motors. We will discuss how they are constructed, their operation, ratings, types and characteristics.

INFORMATION

Construction

Figure 1 shows the two major parts of an AC motor:

1. rotor (rotating part)
2. stator (stationary part)

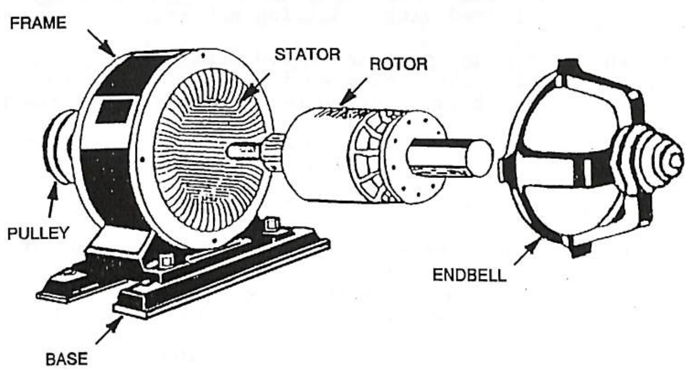


Figure 1.

The other parts shown in Figure 1 include the:

3. frame
4. pulley
5. endbell

Rotor. May be an electromagnet (has current flow) or a permanent magnet (no current flow).

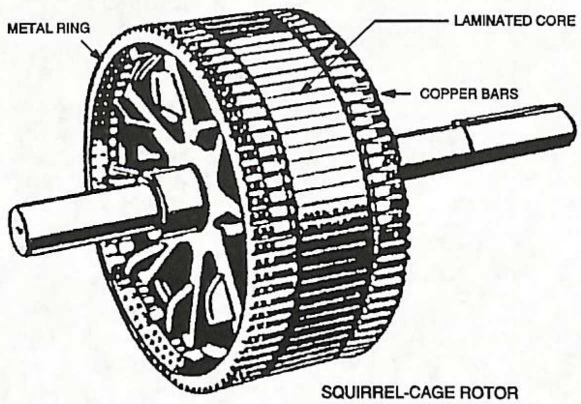


Figure 2.

There are four basic parts to the most common type of rotor (squirrel cage rotor - Figure 2) used with Induction Motors:

a. core which is made up of laminated steel
b. shaft onto which the laminated iron core is placed
c. number of copper bars inserted into slots cut into the laminated iron core
d. ends of copper bars connected to end rings which must be making good contact or the rotor will have power loss

Stator. There are two basic parts to the stator:

a. one, two, or three sets of coils (Determines type of motor)

   1. one set of coils = single phase induction motor
   2. two sets of coils = split phase induction motor/capacitor - start motor
   3. three sets of coils = three - phase induction motor/three - phase synchronous motor

b. laminated steel core with semiclosed slots with coils wound into these

Frame. May be made up of heavy cast iron or steel onto which the core is placed.

Pulley. Attached to the rotor shaft and is the mechanical link to the device driven by the motor.

Endbell. Helps to keep rotor in position so that it will remain centered within the stator and not be allowed to rub on the stator while it is rotating.

Operation of an AC motor

Torque. Torque is the turning force of a motor. An AC motor may consist of a current-carrying conductor formed into a coil and placed on a shaft. The rotor is free to rotate within a revolving magnetic field. The interactions of the magnetic fields, between the rotor and the stator, develop the torque needed to cause the shaft, and thereby, the rotor to turn.

Motor Action. Rotating magnetic fields produce torque. The rotating (changing) magnetic field of the stator must be strong enough to cause mechanical motion of the rotor, to which gears or other mechanical linkages are attached. The stator acts as the primary winding of a transformer, while the rotor acts as the secondary winding. There are no physical connections between the stator and the rotor; all voltages in the rotor are induced.

The principle of induction may be illustrated by placing a horse-shoe magnet over a metallic disk so the magnetic field of the magnet passes through the disk. If the magnet is rotated the disk rotates. Induction motors often use a cylinder rather than a disk and the rotation of the magnetic field is accomplished electrically rather than physically, but the basic principle is the same as that just described.

Single Phase Induction Motor. The one set of stator windings are connected to a single-phase power source which will induce a current into the rotor which produces a constantly changing (revolving) magnetic field. However, some external means (addition of a starter winding or capacitor) must be used to start the rotor moving. The addition of the starter winding or capacitor will be discussed later in this section.

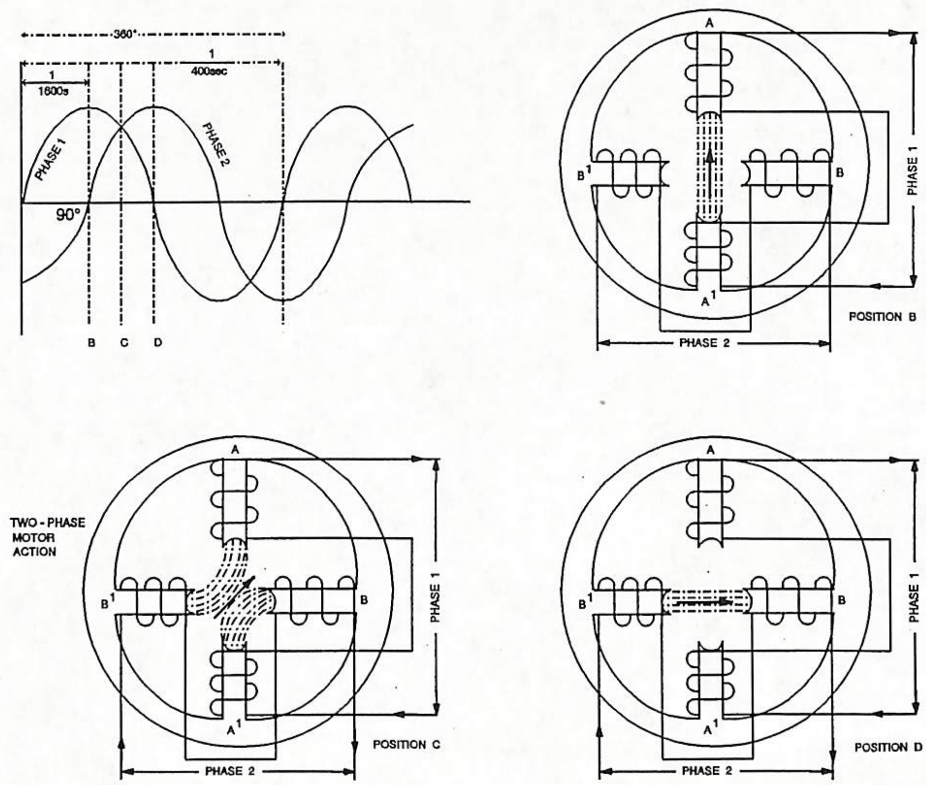


Figure 3.

Although Figure 3 indicates that two sets of stator windings are connected to two separate phases from a two-phase generator, the intent of this example is to show the operating principles of a split-phase motor (split-phase induction motor/capacitor-start motor). In practice, there is one phase from the power source which will become two phases due to placing a starter winding or capacitor into the circuit to cause the currents to be thrown out of phase.

The diagram in Figure 3 shows that Winding 1 (Phase 1) of the motor has a capacitor in the circuit that causes its current to lead the current in Winding 2 (Phase 2) by 90°, assuming a 60Hz AC power supply. The direction of the magnetic field is indicated by a magnetic needle. The needle will line up with the magnetic flux passing from pole to pole.

Phase 1 supplies current to the coils on poles A and A′ and Phase 2 supplies current to the coils on poles B and B′. These currents are 90° out of phase, with Phase 1 leading.

1. At position B, the current in Phase 1 is maximum and the poles of A and A′ are fully magnetized. The poles of coils B and B′ are not magnetized because current in Phase 2 is zero.

2. At position C, the current in the coils A and A′ (Phase 1) has decreased to the same value to which the current in coils B and B′ (Phase 2) has increased. The four poles are equally magnetized, so the strength of the field is concentrated midway between the poles, and the magnetic needle takes the position shown.

3. At position D, Phase 1 current is zero through coils A and A′; these coils are not magnetized. Current is maximum through coils B and B′, the magnetic field strength of B and B′ is maximum, and the magnetic needle takes the horizontal position. This action is repeated during successive cycles of the alternating current. The magnetic needle will continue to rotate in the same direction as long as the two phase currents are supplied to the two sets of coils.

Ratings

AC motors are rated by horsepower, frequency, speed, and voltage.

Horsepower (hp). A unit of power or capacity of a mechanism to do work.

Frequency. The input frequency of AC motors is an important factor in motor operation. As we have seen in our example of the rotating magnetic field, the rate of changing the polarity is what causes the motor to turn properly. If the input frequency does not match the frequency the motor is designed to operate at, the motor speed will be incorrect.

Speed. How fast the rotor shaft is turning. This is expressed in revolutions per minute (RPM). Motor speed is usually synchronized with the input frequency. Therefore, it is important that the input frequency of the motor match the frequency specified by the manufacturer of the motor.

Voltage. This is the specified voltage recommended by the manufacturer.

Types of Motors and Characteristics

Most home and business appliances operate on single-phase AC power, so single-phase AC motors are in widespread use. Induction motors account for a large proportion of these because of rugged construction, maintenance-free operation and low cost.

Single-phase Induction Motor. In a single-phase induction motor, the stator's magnetic field alternates poles as the input sine wave voltage swings from positive to negative from the single-phase power source. The rotor is an electromagnet or a permanent magnet; it rotates to satisfy the attraction created by the constantly changing magnetic field of the stator.

The single-phase induction motor cannot start itself if the rotor and stator are already aligned (a condition called "dead center"). If the rotor and stator are not aligned, when current is applied a small voltage is induced into the rotor, which tends to rotate into alignment. When the rotor reaches dead center voltage is no longer induced, but inertia carries it past dead center and the rotor becomes attracted to the opposite pole because of the change in polarity. Figure 4 shows the polarity characteristics of a single-phase stator winding.

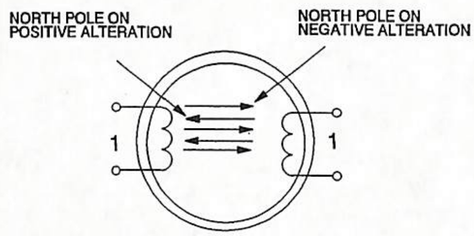


Figure 4.

A single-phase induction motor can be started mechanically but most motors use some type of automatic starting. Single-phase induction motors are classified according to the method used for starting. Two common types include: 1. split-phase motors and 2. capacitor motors.

Split-Phase Induction Motor. Two sets of windings, main winding (run winding) and start winding are included in the stator. The purpose of the two sets of windings is to "split" the single-phase AC voltage into a two-phase current, as shown in Figure 5. The start winding has only a few turns of very light-gauge or thin wire and a small inductance (few turns of wire). The main winding has many turns of large-diameter wire; therefore, the main winding has a smaller resistance (large-diameter wire) and a large inductance (many turns). According to ELI, voltage leads current in an AC inductive circuit. As inductance increases, so does the lag in current flow which would ultimately cause a large phase difference.

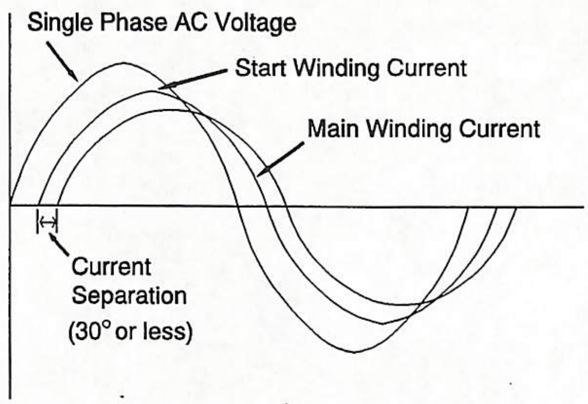


Figure 5.

When current flows through both windings simultaneously, a field builds up much more quickly around the start winding than around the main winding. The current flow in the start winding leads the current flow in the main winding by approximately 30° (Figure 6). The phase difference of 30° results in the motor having a low starting torque. Low starting torque makes these motors ideal for driving mechanical loads, including small machinery and appliances in the home.

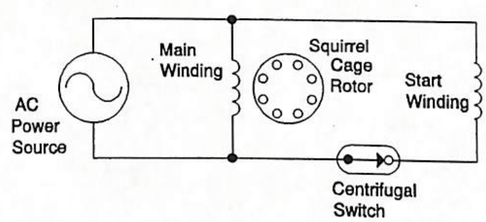


Figure 6.

Once the motor is running almost full speed, the start winding must be removed from the circuit because current through it would burn out the winding. Frequently, a centrifugal switch is inserted so that once it reaches sufficient speed to continue turning, using only the main winding, centrifugal force opens the switch and breaks the continuity through the start winding circuit.

Split-phase Capacitor Motor. Capacitor motors are very similar to the Split-Phase Induction Motors. Like the Split-Phase Induction Motors, the Capacitor Motors have two sets of windings, the main winding and the start winding in the stator. The main difference between the two motors is the addition of 1 capacitor which may be removed in a Capacitor Start Motor via a centrifugal switch. The Split-phase Capacitor Run Motor has no centrifugal switch, and therefore, no means of removing the capacitor.

Capacitor-Start Induction Motors. The Capacitor-Start Motors have 1 capacitor which is in series with the start winding and the centrifugal switch, as shown in Figure 7.

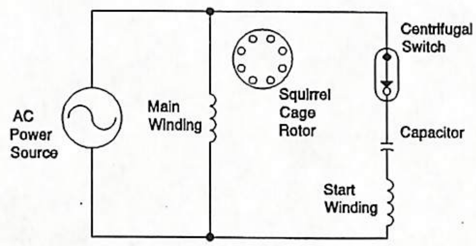


Figure 7.

The capacitor, according to ICE (current leads voltage in a capacitive circuit), is responsible for causing the current in the start winding to lead the single phase AC voltage from the input, as shown in Figure 8. This phase difference may be attributed to the high capacitance in the start-winding portion of the circuit. The main winding's portion of the circuit only has inductance (ELI) so its current will lag the AC voltage from the input.

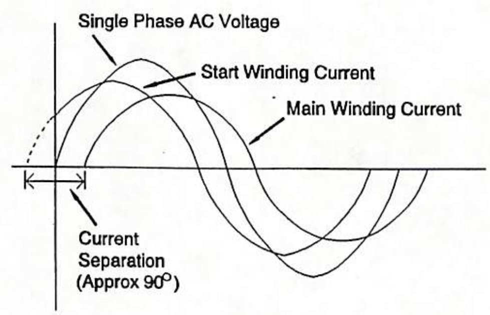


Figure 8.

The phase difference between the two currents is very close to 90°; therefore, this type of motor may be used where high starting torque is essential. When the rotor reaches sufficient speed the start winding is switched off by the centrifugal switch, and the motor continues running on the main winding.

Split-phase Capacitor-run Motor. This particular capacitor motor differs from the one previously discussed in that it has NO centrifugal switch. The capacitor remains in the circuit at all times. When very low torque is necessary these motors are used. Figure 9 indicates the circuit for the Split-phase Capacitor-run Motor.

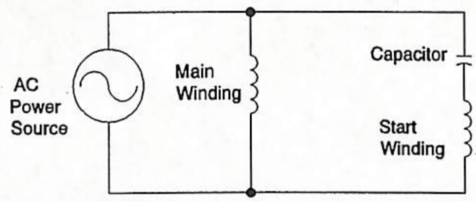


Figure 9.

Split phase-Induction Motors and Capacitor Motors may need to have their direction of rotation reversed. The start winding connections OR the Main windings connections must be reversed to reverse the direction of rotation. DO NOT REVERSE BOTH!!

Synchronous Motor. Three Phase AC Synchronous Motors are very specialized motors. (Figure 10) These motors have the characteristic of constant speed between no load and full load. They are capable of correcting the low power factor of an inductive load when they are operated under certain conditions.

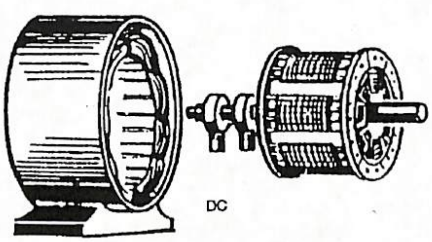


Figure 10.

To understand how the synchronous motor works, assume that the application of three-phase AC to the stator causes a rotating magnetic field to be set up around the rotor. The rotor's wire windings are connected by slip-rings and brushes to a DC power source. The strong rotating magnetic field from the stator attracts the strong rotor field activated by the DC. This results in a strong turning force on the rotor shaft. The rotor is therefore able to turn a load as it rotates in step with the rotating magnetic field.

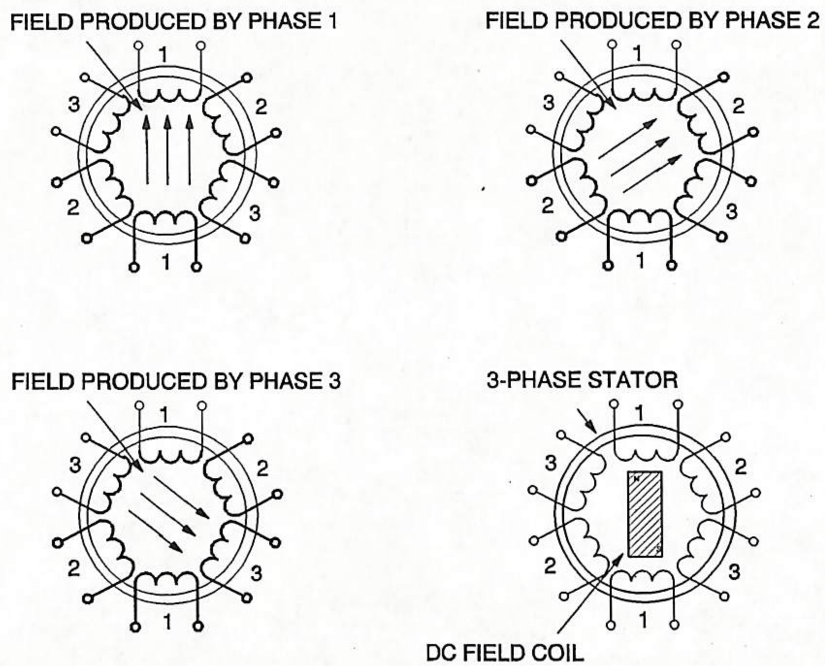


Figure 11.

Figure 11 shows a three-phase AC synchronous AC motor. It works this way once it's started. However, one of the disadvantages of a synchronous motor is that it cannot be started from a standstill by applying three-phase AC to the stator. When AC is applied to the stator, a high-speed rotating magnetic field appears immediately. This rotating field rushes past the rotor poles so quickly that the rotor does not have a chance to get started. In effect, the rotor is repelled first in one direction and then the other. A synchronous motor in its purest form has no starting torque. It has torque only when it is running at synchronous speed. These motors are designed such that they will rotate at a constant speed regardless of the load. This constant speed is called synchronous speed. The synchronous speed of a synchronous motor is based on the speed formula:

$$\text{speed (rpm)} = \frac{\text{AC frequency} \times 120}{\text{number of poles} / 3}$$

Synchronous AC motors have speeds that are based on the number of stator poles and the AC frequency.

A squirrel-cage type winding may be added to the rotor of a synchronous motor to cause it to start. The squirrel-cage is shown as the outer part of the rotor in Figure 12. Simply, the windings are heavy copper bars shorted together by copper rings. A low voltage is induced in these shorted windings by the rotating three-phase stator fields. Because of the short circuit, a relatively large current flows in the squirrel-cage. This causes a magnetic field which interacts with the rotating field of the stator. Because of the interaction, the rotor begins to turn, following the stator field; the motor starts.

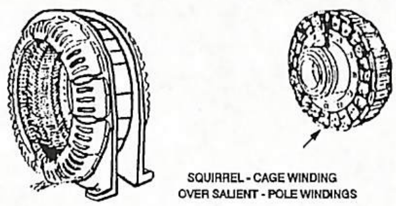


Figure 12.

To start a practical synchronous motor, the stator is energized, but the DC supply to the rotor field is not energized. The squirrel-cage windings bring the rotor field to near synchronous speed. At that point, the AC field is energized. This locks the rotor in step with the rotating stator fields. Full torque is developed, and the load is driven. A mechanical switching device that operates on centrifugal force is often used to apply DC to the rotor as a synchronous speed is reached.

Universal Motor. A universal motor operates on either AC or DC, as shown in Figure 13. It is constructed as a series-wound DC motor. This arrangement is the only type which will operate with AC applied. Series wound DC motors have windings with low inductance due to having few turns of large-diameter wire. Therefore, these windings offer a small resistance to alternating current flow. The universal motor has windings similar to those of series wound DC motors.

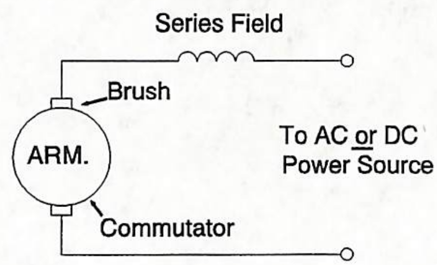


Figure 13.

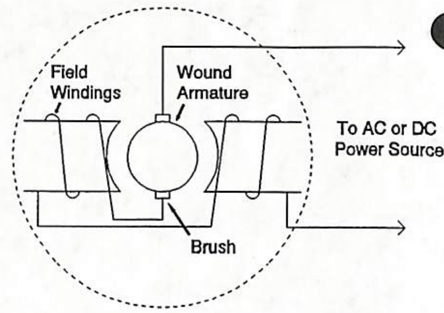


Figure 14.

The operation of the universal motor with alternating current applied causes a change in field and armature polarities at the same time (refer to Figure 14). Alternating current brings about a reversible field and armature polarity. Reversing the direction of current flow through the armature windings at the proper time causes the armature to rotate. The universal motor operates in the same way as the series-wound DC motor except that the field polarity and the direction of the current through the armature windings changes at a rate of 120 times/second with 60Hz AC applied. Universal motors speed and torque are similar to those of series-wound DC motors. Some applications of universal motors include: mixers, blenders, drills, and saws.

SUMMARY

In this lesson we covered the purpose, construction, torque, operation, ratings, and types and characteristics of AC motors.

13.2.5 Synchro and Servo Operation

SYNCHRO/SERVO OPERATION

OBJECTIVES

Identify synchro system operating principles.

Identify servo system operating principles.

INTRODUCTION

Automatic control systems currently used in industry and the military provide precise control of angular positions. Synchros and servos are commonly used together in one automatic control system. Precision control requires indicating the exact position of things like control surfaces (rudders, ailerons, flaps, etc.), antennas, and gun turrets as well as accurately positioning them. Generally, the devices that make up the indicating portion are synchros, while devices that make up the positioning portion are servos.

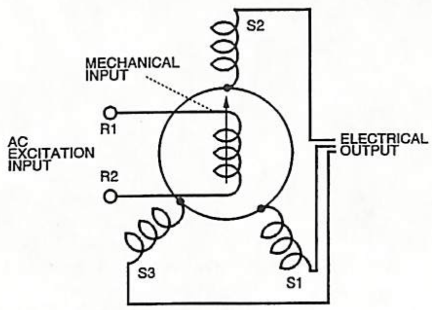
Basically, synchros are light-duty components that use low-current, low-power control, error, and indicating signals, while servos are heavy-duty components that use high-current, high-power signals to position loads. Most automatic control systems also contain error detectors and servo amplifiers. Error detectors detect any difference between where a load is commanded to be and its actual position. Servo amplifiers convert the low-level control and error signals into high-level positioning signals.

INFORMATION

Synchro Schematic Symbols

Symbols and Labeling. So you can better understand what the schematics represent electrically, first look at electrical representations of synchros.

Basic synchro. The electrical representations of basic synchros (Figures 1 and 2) are similar to those of three-phase AC generators. They also look a little like a "wye" (Y) transformer: they have three windings spaced 120° apart, and they also have a movable coil in the center. The movable coil, or rotor, can function as either the primary or secondary winding in relation to the wye portion, or stator, depending on how the synchro is used. The rotor serves as the primary, with the stator as the secondary, when the synchro is used as a synchro-generator (G) or synchro transmitter (TX). The stator serves as the primary, with the rotor as the secondary, when the synchro is used as a synchro-motor (M) or synchro receiver (TR). In either case, the 0°, or "null," position occurs when the rotor winding's axis is directly in line with stator winding 2 (S2).



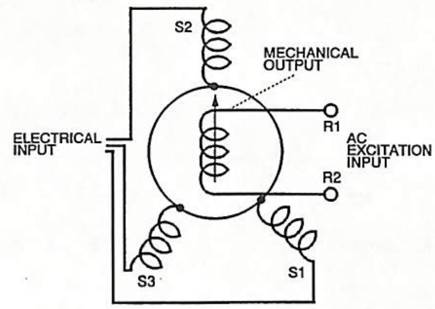
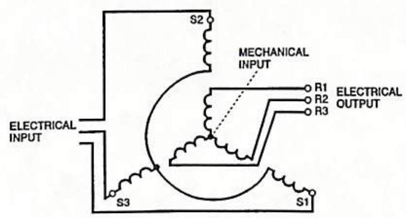


Figure 1. Synchro Generator

Figure 2. Synchro Motor

Differential synchro. Differential synchro electrical diagrams (Figures 3 and 4) are almost the same as basic synchros except that the rotor also has three parallel windings spaced 120° apart. When a differential synchro is used as a torque-differential synchro-generator, or torque-differential transmitter (TDX), the rotor is the primary. When it is used as a torque-differential synchro-motor, or torque-differential receiver (TDR), the rotor is the secondary. The null position occurs when rotor winding 1 (R1) is directly in line with S2.



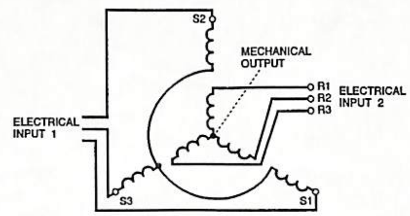


Figure 3. Differential Synchro-Generator (TDX)

Figure 4. Differential Synchro-Motor (TDR)

Control transformer. A control transformer (CT) electrical representation (Figure 5) is very similar to that of a basic synchro, except that the null position occurs when the rotor winding's axis is at a 90° angle in relation to S2.

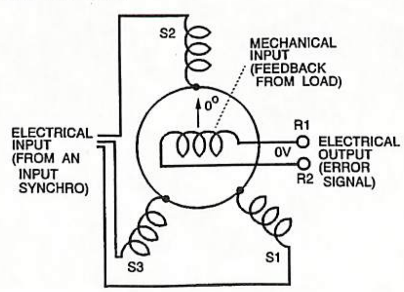


Figure 5. Control Transformer

Just as the electrical representations for synchros are similar, so are the schematic diagrams (Figures 6 and 7). The synchro-generator and synchro-motor schematics are identical except that one input and the output are reversed. One input, the AC excitation voltage, is applied to the rotor in both cases. The other input to the synchro-generator is mechanical movement of the rotor (for example, by a control dial); the output is an electrical change in the phase of the signals at the stator windings. The other input to the synchro-motor is an electrical change of the phase in the signals at the stator windings while the output is mechanical movement of the rotor (for example, movement of an indicating needle).

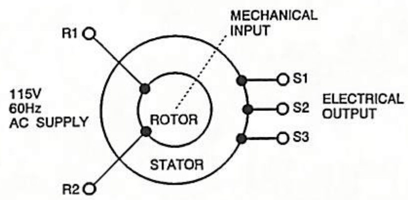


Figure 6. Synchro-Generator Schematic

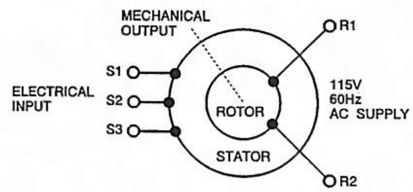


Figure 7. Synchro-Motor Schematic

A similar situation exists for differential synchro schematics (Figures 8 and 9). The inputs to a differential synchro-generator are three-phase signals at the stator windings and a mechanical movement of the rotor; the output is electrical change in the phase of the signals at the rotor windings. The inputs to a differential synchro-motor are three signals of different phase on both the stator and rotor windings while the output is mechanical movement of the rotor.

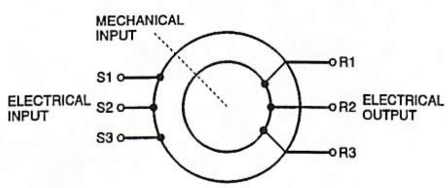


Figure 8. Differential Synchro-Generator (TDX) Schematic

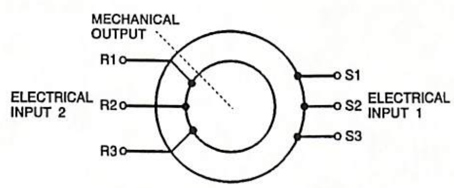


Figure 9. Differential Synchro-Motor (TDR) Schematic

The schematic diagram of a control transformer (Figure 10) resembles those of the synchro-generator and synchro-motor. The inputs to a CT are the three-phase electrical signals on the stator windings and the mechanical position of the rotor; the output is an electrical "error" signal on the rotor windings. The mechanical input, called "feedback", usually is provided by a linkage from the load; it represents the position of the load relative to a fixed point. Any difference between where the load is supposed to be (electrical input) and where it actually is (mechanical input) results in an error signal. When the load has moved to where it is supposed to be, the electrical signals of the rotor and stator agree and no error signal is generated.

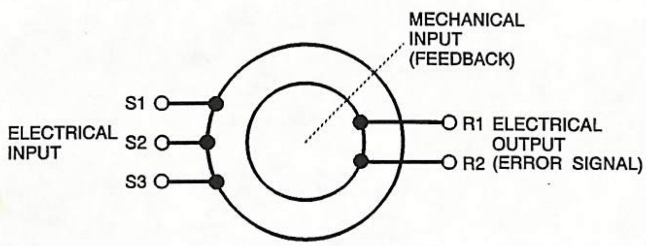


Figure 10. Control Transformer Schematic

Synchro System Operating Principles

Purpose of Synchro Systems. A synchro is the part of an automatic control system used for low-power, remotely-controlled, shaft repositioning/indicating applications. It is an inductive device that transforms an angular position input to an electrical output, or an electrical input to an angular position output. Angular position refers to the degree of rotation of a shaft.

A synchro couples the rotational position of some other component, like a radar antenna, to an indicating device without using a mechanical linkage. The instantaneous "azimuth" position of a radar antenna is shown on a radar indicator by the displacement of an illuminated line. Azimuth is the horizontal, or left/right, scanning movement of an antenna away from its center, or "boresight", position. The angular position of the illuminated line is <u>synchronized</u> with the angular position of the antenna through the use of a synchro-generator and synchro-motor. As the antenna rotates, or scans, the angular position of the illuminated line on the indicator changes along with the antenna. The synchro-generator and synchro-motor form the basic synchro system, although many other devices may be included in the system. A similar set-up is used in radio navigation equipment, such as tactical air navigation (TACAN). The position of a compass needle is synchronized with the TACAN antenna indication.

Construction. A synchro system requires, as a minimum, a synchro generator and a synchro motor. These devices resemble small electric generators/motors: each has a stationary element (the stator) and a movable element (the rotor).

Stator. The stator (Figure 11) consists of three coils placed in slots around the inside of a laminated-iron field structure, spaced 120° apart as in a three-phase generator or motor. The coils overlap slightly and the amount of force, either pulling the rotor into position or sensing the rotor position, is equal, independent of the position of the rotor. In other words, the rotor is attracted to the 0° position from near zero with the same amount of force as if further away from zero. This force is computed using complex mathematical operations beyond the scope of this course.

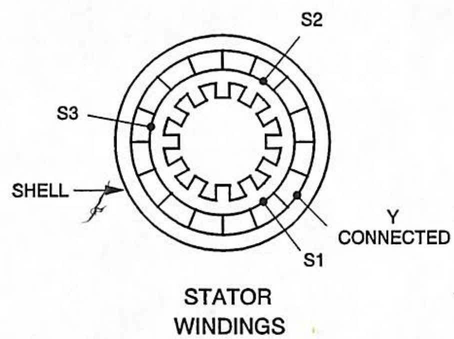


Figure 11. Stator Front View

Rotor windings. The rotor (Figure 12) consists of a coil, or coils, wound on a soft-iron core and mounted on a shaft so that the axis of the coil is perpendicular to the shaft. This causes maximum inductive interaction between the magnetic fields developed on the rotor winding(s) and the stator windings. Ball bearings reduce the friction between the rotor shaft and its mount. The rotor is electrically connected to the external power source through two slip rings on the shaft.

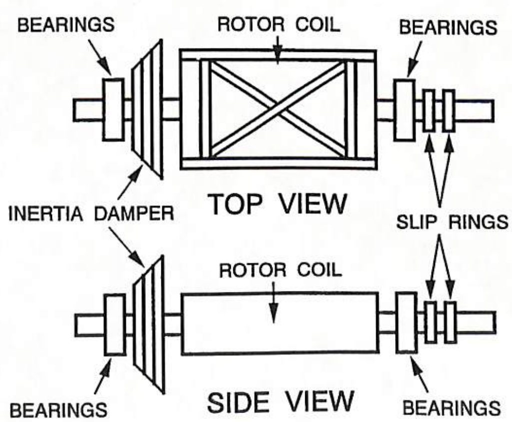


Figure 12. Rotor

Inertia damper. Synchro motors have a tendency to oscillate violently, or spin continuously, when power is first applied to the system or when the shaft is turned suddenly. "Damping" is required to prevent this. Damping is the controlled application of friction to oppose rapid changes in shaft position. The synchro motor contains a heavy flywheel, the inertia damper (Figures 12 and 13), mounted on one end of the shaft so that it turns freely for approximately 45° and then strikes a keyed bushing (Figure 13). The bushing is fastened to the shaft by a friction disc, which also turns on the shaft but with a great deal of friction. The flywheel easily follows slow changes in shaft position. If the shaft moves suddenly, the flywheel tends to stand still because of at-rest inertia and the friction disc acts as a brake, keeping the shaft at a speed too slow to start oscillation or free-spinning. If oscillation or spinning does occur, something is wrong with the damper.

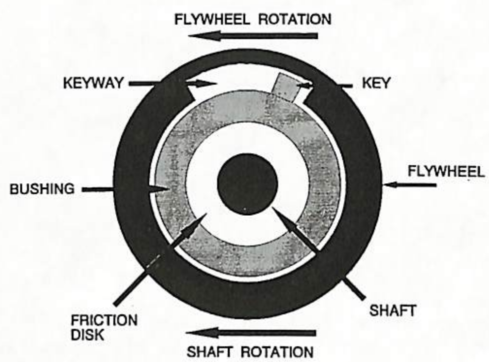


Figure 13. Inertia Damper

Types of Synchros. The four types of synchros are generator, motor, differential, and control transformer.

Generator. In the synchro-generator, the AC excitation voltage (applied to the rotor winding) is coupled to the stator windings through electromagnetic induction. The phase and amplitude of the voltage induced in each stator winding is determined by the angular position of the rotor; both rotor and stator may be stationary. AC current through the rotor causes an electromagnetic field to expand and collapse at the AC rate, cutting across the stator windings and inducing AC voltage and current into them. The amount of voltage induced in a stator winding depends on the angular relationship between the individual stator winding and the rotor coil(s). When two coils are parallel to each other (either end-to-end or in-line), coupling between them is maximum; when two coils are perpendicular to each other, coupling is minimum.

Figure 14 illustrates the effect of angular relationship on coupling. The rotor winding has 115 VAC (60Hz) applied and is parallel to S2, so the magnetic field of the rotor induces 52 VAC (maximum) into S2 (Figure 14A). When perpendicular (Figure 14B), it induces 0 VAC (minimum). When parallel but 180° out of phase, it induces 52 VAC (maximum) of the opposite polarity (Figure 14C).

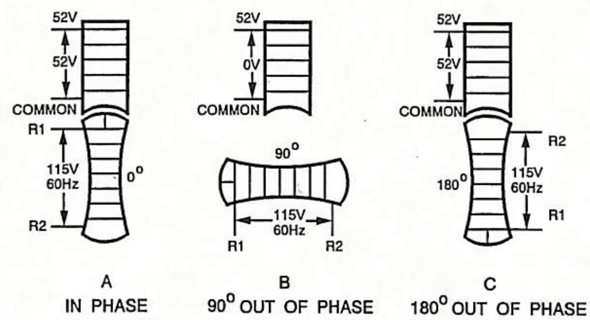
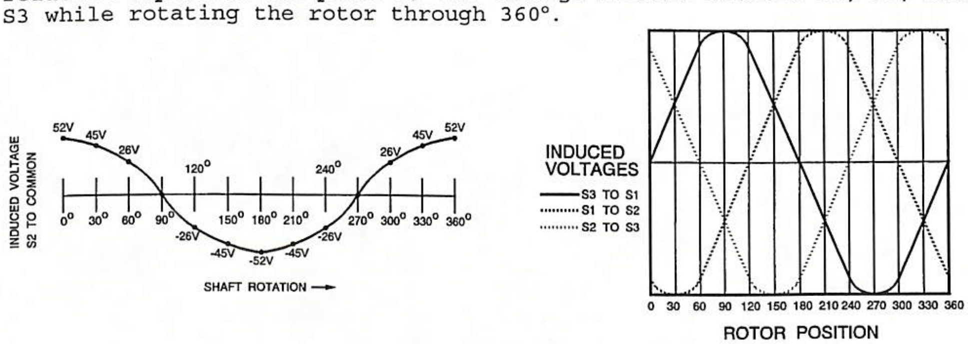


Figure 14. Angular Relationship

As the rotor in a synchro is turned through 360°, the voltage induced in each stator varies. Figure 15A shows the amplitude and phase of the induced voltage for one stator winding (S2). Values plotted above zero indicate an in-phase condition between rotor and stator, while values plotted below zero indicate a 180° out-of-phase condition. The curve is a sine wave of a frequency determined by the time it takes the rotor to turn through 360°. No two stators can be at the same points on their curves simultaneously because they lie 120° apart. Figure 15B shows the relative amplitude and phase of the voltage induced between S1, S2, and S3 while rotating the rotor through 360°.



A B

Figure 15. Induced Voltages (Angular Position)

Each stator reaches maximum induced voltage twice in a 360° rotation of the rotor. The reference line represents the point at which the polarity changes, when the induced voltage is zero. This change of phase in the stators, with respect to the phase of current in the rotor, is shown in Figure 16.

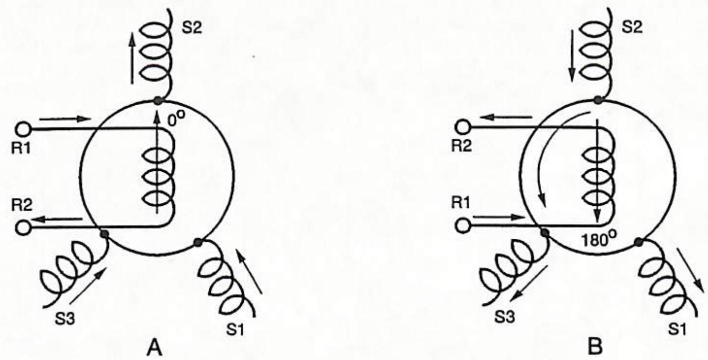


Figure 16. Phasing

NOTE: In standard synchros, the maximum voltage induced in each stator coil is 52VAC when 115VAC is applied to the rotor.

The angular position of the synchro rotor determines the amplitude and phase of the voltage induced in the stator windings. In a synchro generator, the angular position of the rotor is converted into electrical signals of different phase by the stator windings. The output of the synchro-generator may then be coupled to a synchro-motor. In the synchro-motor, the electrical signals on the stator windings develop magnetic fields that position the rotor. When the magnetic field around S2 is maximum and is developed in the same direction as the magnetic field around the rotor, the rotor lines up with S2, much like a bar magnet placed in the center of three bar magnets spaced 120° apart. Figure 17 shows the connections between the synchro-generator and synchro-motor.

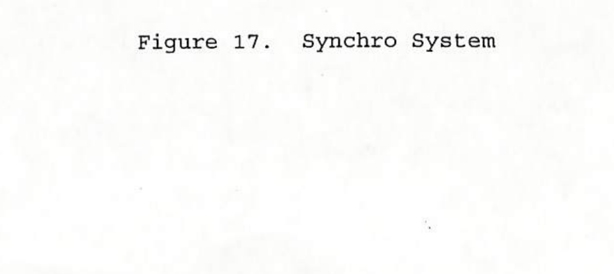


Figure 17. Synchro System

Motor. A synchro-motor also can be positioned by an input from an autotransformer that simulates the output from a synchro-generator (Figure 18). The arrows indicate the direction of current flow in rotor and stators for the positive half-cycle of applied voltage. S1 and S3 have equal magnetic field strength and polarity. S2 has the greatest magnetic field strength as all current that flows through S1 and S3 must flow through S2. The polarities of the magnetic fields lock the rotor in the position shown. On the negative half-cycle, all currents in the synchro are reversed so the polarities of the rotor and stator magnetic fields reverse. The rotor remains locked in the position shown and the net torque on the rotor is zero.

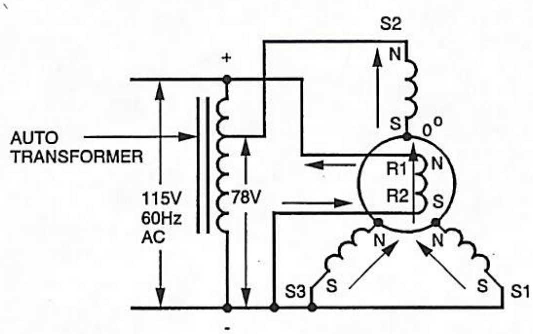


Figure 18. Synchro-Motor with Autotransformer Input

The current phase in the stator windings in Figure 19A has been reversed from that of Figure 18 by switching the leads from the autotransformer to the stators. The rotor has moved to the 180° position to align the rotor and the stator magnetic fields. Two positions of the rotor can be obtained by shifting the current phase in the three stator windings with respect to the current phase in the rotor windings. To position the rotor anywhere other than 0° and 180°, either the relative amplitude or phase of current in the individual stators must be varied. Stator S3 in Figure 19B has been disconnected. The rotor now aligns with the magnetic fields that exist between S2 and S1.

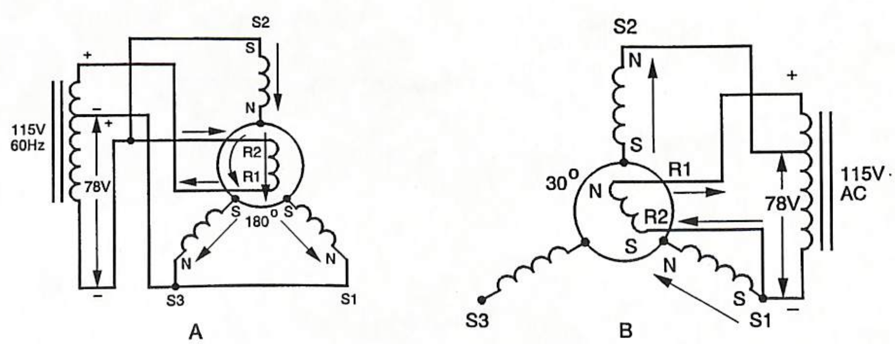


Figure 19. Stator Winding Inputs

Differential. There are two types of torque-differential units, differential generators (transmitters) and differential motors (receivers). The torque-differential transmitter (TDX) accepts one electrical input and one mechanical input and produces one electrical output. The torque-differential receiver (TDR) accepts two electrical inputs and produces one mechanical output. The torque-differential transmitter or the torque-differential receiver can be used to form a differential synchro system. A TX-TDX-TR system consists of a synchro-transmitter (TX), a torque-differential transmitter (TDX), and a synchro-receiver (TR); a TX-TDR-TX system has two synchro-transmitters (TXs) and one torque-differential receiver (TDR).

In the torque-differential transmitter, both the rotor and the stator windings consist of three wye-connected coils (Figure 20). Usually the stator is the primary and receives its input signal from a synchro-transmitter. The voltages appearing across the rotor terminals (R1, R2, and R3) are determined by the strength of the magnetic fields, the position of the rotor, and the turns ratio between stator and rotor windings. The magnetic fields created by the stator currents assume an angle corresponding to that of the magnetic field in the transmitter supplying the signal.

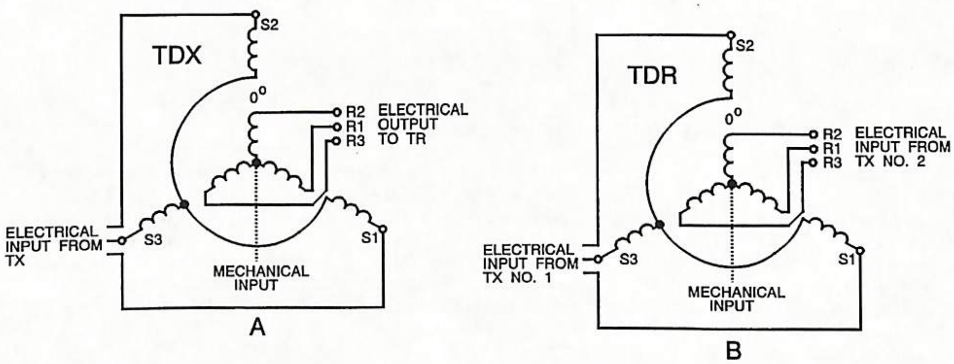


Figure 20. Torque-Differential Transmitter and Receiver

The position of the rotor controls the amount of magnetic coupling (between the stator magnetic field and the rotor) and the amount of voltage induced into the rotor windings. If rotor position changes in response to a mechanical input, the voltages induced into the windings also change. The TDX output voltage varies as a result of a change in the input stator voltage or a change in the mechanical input to the rotor. The TDX electrical output may be either the sum or the difference between the two inputs, depending on how the three units (TX, TDX, and TR) are connected.

The TDX and the TDR are electrically identical. The TDR also has a damper to prevent the rotor from oscillating. The receiver provides the mechanical output for a differential synchro system: the sum or difference of two electrical inputs from synchro-transmitters. As with the TDX, the TDR addition or subtraction is a function of how the units in the system are connected. The TDR operates like an electromagnet. In Figure 20B, the TDR rotor and stator receive energizing currents from two TXs that produce two resultant magnetic fields, one in the rotor and the other in the stator. Each magnetic field assumes an angle corresponding to that of the magnetic field in the transmitter. The interaction of these two resultant magnetic fields causes the TDR rotor to turn.

Control transformer. A control transformer is a synchro device that accurately governs a power amplifier to move heavy equipment. A CT is similar to a synchro-generator or synchro-motor except that there is no damper and the rotor is a drum or wound rotor rather that an elliptical-pole rotor. The CT compares the electrical signal at the stator with the mechanical signal at the rotor. The CT output is a difference signal (error signal) that controls a power-amplifying device.

The CT rotor is never connected to an AC supply and it does not induce voltages in the stator coils. CT stator currents are determined solely by the voltages applied to the high-impedance stator windings; rotor position has little effect. No appreciable current flows in the rotor because its output voltage is applied to a high-impedance load. The CT rotor does not try to follow the magnetic field of its stator and must be turned by some external force (usually a mechanical feedback linkage).

The CT stator windings are the primary windings and the rotor winding is the secondary winding. The output (taken from the R1 and R2 rotor leads) is the voltage induced in the rotor winding. Phase and amplitude of the output voltage depend on the angular position of the rotor with respect to the magnetic fields of the stator windings.

Synchro System Operation. Synchros work in teams. Two or more synchros are interconnected electrically to form a synchro system. A generator and motor, properly connected, form a complete synchro system in which the motor shaft follows the generator shaft (Figure 21).

Connections. Both R1 terminals are connected to one side of the supply line and both R2 terminals to the other so that the two rotors have voltages applied that are identical in magnitude, phase, and frequency.

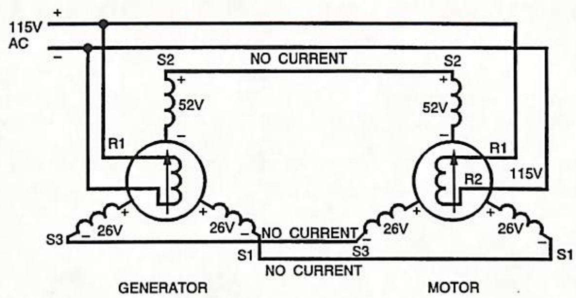


Figure 21. Basic Synchro System

Each motor lead is connected to its corresponding generator lead. Each stator winding of the motor has a voltage induced in it that is opposed by an equal voltage in the corresponding stator winding of the generator. The conditions in Figure 21 exist when motor and generator rotors are at the 0° position. No current flows in the stator windings of either motor or generator. When the rotor positions change, the voltages induced in the stators change, but if both motor and generator are in corresponding positions, no current flows in the stators and no torque is produced.

Electrical zero. To work together properly in a system, synchros must be connected correctly and aligned with each other and with the load devices. Electrical zero is the alignment reference point of all synchro units. The mechanical reference point for the units connected to the synchros depends upon the particular application of the system. The mechanical position is set first, then the synchro device is aligned to electrical zero; the system is in its null position and no electrical difference exists between generator and motor.

Current flow. When the rotors are not in corresponding positions, the voltage induced in corresponding stators is not equal and current flows in the stator windings of both motor and generator (Figure 22). The current flowing in the stators causes a magnetic field to interact with the magnetic field around the rotors, exerting a torque on the rotors of both motor and generator. The torques in the motor and generator are opposite in direction; both try to turn the rotors toward corresponding positions. In normal synchro systems, the generator rotor is positioned by an external device. Only the motor rotor is free to turn toward a position of correspondence. Stator currents are almost directly proportional to the difference in rotor positions.

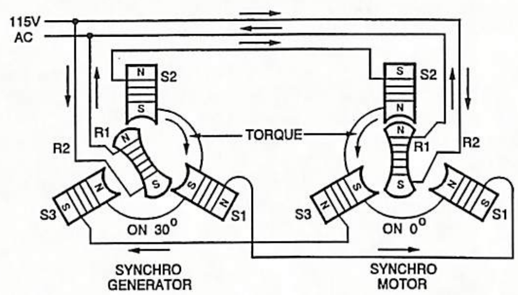


Figure 22. Stator Windings Not Aligned

Error signal. To produce any change, a synchro system must respond to an error signal. The demands on a synchro system may be more complex than positioning an indicator in response to information received from a single signal source. An error detector for checking weapons uses a synchro system to determine the error in gun position with respect to the positioning order. Such a synchro system must accept two signals, one containing the positioning order and the other corresponding to the actual position of the gun. The system must compare the two signals and position an indicator to show the difference.

INFORMATION

Servo System Operating Principles

Purpose of Servo Systems. Servo systems have many applications in electrical and electronic equipment operation. In radar and antennas, directors, computing devices, ship's communications, aircraft control, and other equipment, it is often necessary to operate a mechanical load that is remote from its source of control. For smooth, continuous, and accurate operation, these loads are controlled by synchros. Synchros alone are not powerful enough to do any great amount of work.

Definition. A servo system uses weak control signals to move large loads to a desired position with great accuracy. Servos are found in varied applications such as moving the rudder and elevators of an airplane or controlling the diving planes and rudders of nuclear submarines. Servos are powerful; they can move heavy loads and are remotely controlled with great precision by synchro devices.

Forms of servos. Servo systems may be either electromechanical, electrohydraulic, hydraulic, or pneumatic. Whatever the form, a relatively weak signal representing a desired movement of the load is generated, controlled, amplified, and fed to a servo motor that does the work of moving the heavy load.

Types of Servos. Servo systems are a part of a broad category of control systems: groups of components linked together to perform a specific function. Generally, a control system has a large power gain between input and output. The components used in the system and the complexity of the system are directly related to the application requirements. Control systems are broadly classified as either <u>closed-loop</u> or <u>open-loop</u> systems.

Open-loop system. An open-loop control system is controlled directly, and by an input signal only. The basic units of the system are an amplifier and a motor. The amplifier receives a low-level input signal and amplifies it to drive the motor to perform the desired task. Open-loop control systems are not used as commonly as closed-loop control systems because they are less accurate. Closed-loop control systems are frequently used in military applications because they respond quickly and move the loads they are controlling with greater accuracy than open-loop systems.

The quicker response and greater accuracy is due to an automatic feedback system that confirms the desired movement has taken place. Upon receiving this information, the system stops the motor; load motion ceases until another movement is ordered by the input. This is similar to the thermostat system that controls heat in many homes: the thermostatic input calls for heat and the furnace produces heat output and distributes it. Some of the heat is "fed back" to the thermostat. When this feedback raises the temperature of the thermostat to that of the setting, the thermostat responds by shutting the system down until heat is again required. In such a system, the feedback path (input to output and back to input) forms a closed loop. Closed-loop control systems are automatic in nature, so they are classified by the function they serve (controlling position, velocity, or acceleration of the load).

In an open-loop system, the behavior of position of the output shaft is not matched automatically with the input position. Matching or correction is accomplished manually. An example of an open-loop servo system is a steering system used in a ship (see Figure 23). The rudder turns in a direction and an amount determined by the rotation of a wheel in the pilot house. The rudder is large and heavy, so it is pivoted by a servomotor.

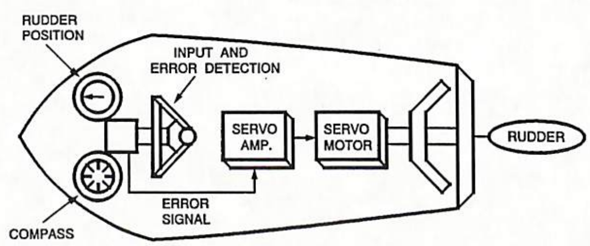


Figure 23. Open-Loop Servo System

The helmsman serves as both the input and the error detector, moving the wheel to keep the ship on course. The helmsman watches the heading on a compass; when the wind or ocean current moves the ship off course, the helmsman turns the wheel (and consequently the rudder) to bring the ship back on course. This is an open system because there is no feedback from the output function, and it is not automatic.

Closed-loop system. In a closed-loop servo system, input and feedback signals are automatically compared to bring the input and output conditions into automatic alignment. If the helmsman is replaced with an electronic device (an automatic pilot), the steering system becomes a close-loop system (see Figure 24).

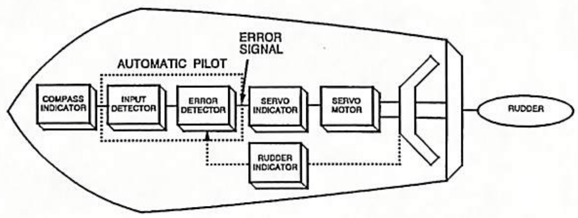


Figure 24. Closed-Loop Servo System

When the ship drifts off course, the change in heading generates a voltage signal that is transmitted to the error detector. The rudder and heading (compass) positions no longer agree, so an error signal is sent through the servo amplifier to the motor to turn the rudder a specific amount in the direction required to bring the ship back on course. By the time that the ship is back on course, the rudder will have returned to the amidships (straight ahead) position.

For example, if a ship is heading west (270° compass heading) and the wind pushes it to a heading of 265°, there is a five-degree angular difference between the established rudder and compass dial positions (see Figure 25).

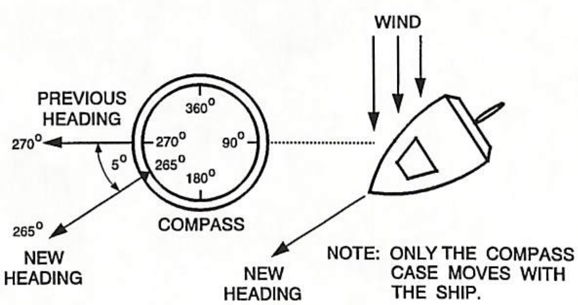


Figure 25. Reading/Heading Comparison

When the compass casing moves to the 265° position, it signals to the autopilot that the compass components (dial and casing) no longer agree; the autopilot generates an error signal of the amplitude required to compensate for the deviation. The rudder is turned to the right five degrees to bring the ship back on course.

As the ship returns to 270°, the rudder angle is decreased correspondingly and it returns to center when the ship heading again becomes due west. Now the compass (dial) and rudder positions are the same and the error signal is zero. Feedback from the output function and automatic comparison characterize a closed-loop servosystem.

Operation. The servosystem shown in Figure 26 is the most common type today.

Basic system. It is made up of electromechanical parts and consists of a synchro control system, servo amplifier, servo motor, and some form of feedback (response). The synchro control system controls the movement of the load, which may be located in a remote area; the servo amplifier and servo motor actually develop the power to move the load because the controlling signal from a CT is too weak to drive an electric motor directly.

In Figure 26A, the control signal is initiated by a handcrank input connected to the synchro transmitter (TX). The dials located on the TX and the CT indicate the positions of the synchro's rotors, while the dial on the load indicates the position of the load. In Figures 26A and 26D, the dials of the TX and load indicate that the load is in the desired position; so there is no error signal present at the servo amplifier and no power to the servo motor.

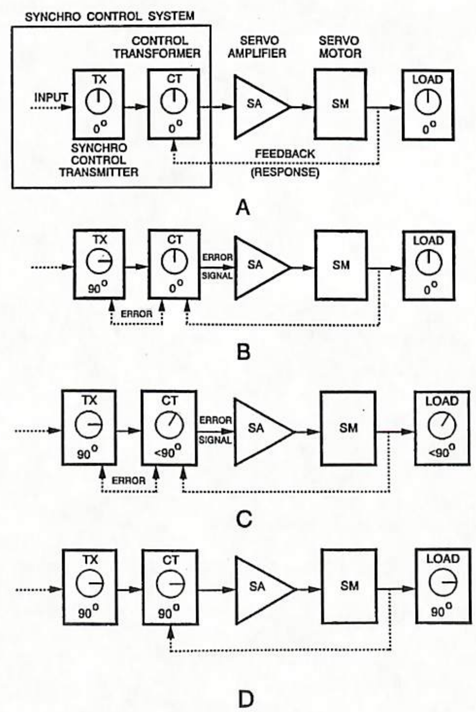


Figure 26. A Closed-Loop Servo System

Control Transformer (CT). In Figure 26B, the TX rotor has been moved by the handcrank to 90°, indicating that it is to move the load 90°. The CT rotor remains at 0°. The CT develops an error signal proportional in amplitude to the difference in position between the CT rotor and the TX rotor. The phase of the error signal indicates the direction the CT rotor must move to reduce the error signal to zero, or null it out. The error signal is sent to the servo amplifier. In Figure 26C, the error signal, amplified by the servo amplifier, is sent to the servo motor.

Servo motor. The motor starts to drive in the direction that will reduce the error signal and bring the CT rotor back to the point of correspondence, in this case clockwise. The mechanical linkage attached to the servo motor moves the CT rotor also. This feedback causes the error signal amplitude to decrease, slowing the speed at which the load is moving.

Connections. In Figure 26D, the servo motor has driven both the load and CT rotor so that the CT and the TX rotor positions correspond; the error signal is reduced to zero and the load stops at its new position. This servo system moved a heavy load to a predetermined position in response to the turning of a handcrank.

Two key points about the operation of the closed-loop servo system are:

1. The original error (movement of the TX rotor) was "detected" by the CT--the error detector.

2. The servo motor, in addition to moving the load, provides mechanical feedback to the CT to reduce the error signal. The servo motor is an error reducer.

Positioning. Figure 27 shows the basic operation of a typical position-changing servo that has wide application in military equipment. An input order indicates a position in which the load--a gun turret--is to be placed. The load in Figure 27 is a gun turret.

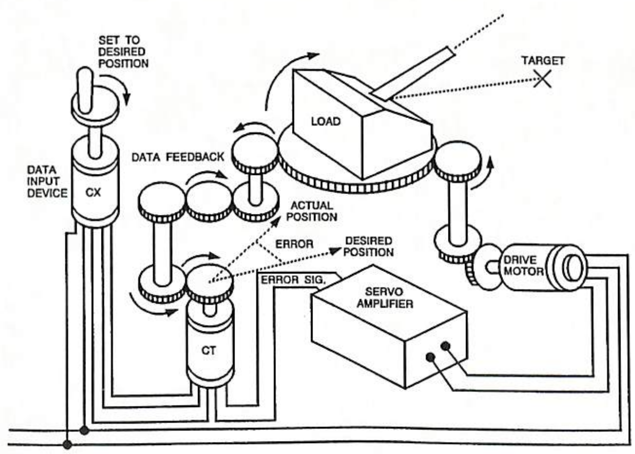


Figure 27. Typical Position Servo

The purpose of the system is to position the gun by means of an order from a remote handcrank. The load is mechanically coupled through a gear train to the rotor of a CT so that turret position is always accurately represented by CT rotor position. An order signaling the desired position of the gun turret is fed into the servo by positioning the TX rotor with the handcrank. A corresponding signal immediately appears across the CT stator. This signal differs from the actual position of the gun turret, causing an error voltage to be developed across the CT rotor.

Error signal. The error voltage is fed from the CT rotor to the servo amplifier, which converts it into power with a polarity or phase relationship to drive the motor in the direction necessary to bring the gun into the desired position.

Feedback. As the turret moves, mechanical feedback turns the CT rotor towards agreement with the TX rotor. As the load approaches the proper position, less and less power is supplied to the motor because of the decreasing error voltage developed in the CT. When the electrical position of the CT rotor agrees with the position of the TX rotor, the error voltage reaches zero and power is removed from the motor. The turret is now in the desired position.

Typical position servo. In an actual system (see Figure 28), the heavy gun turret's momentum tends to carry it past the desired position. This overshoot causes the rotor of the CT to move out of correspondence with the TX rotor. A new error signal develops, opposite in polarity to the original input signal. The new error signal opposite direction. The correcting process, called <u>hunting</u>, may continue until the gun is positioned correctly, with each succeeding correction slightly smaller than before. In most servos an antihunt electronic network or damping system is used to minimize this undesirable effect.

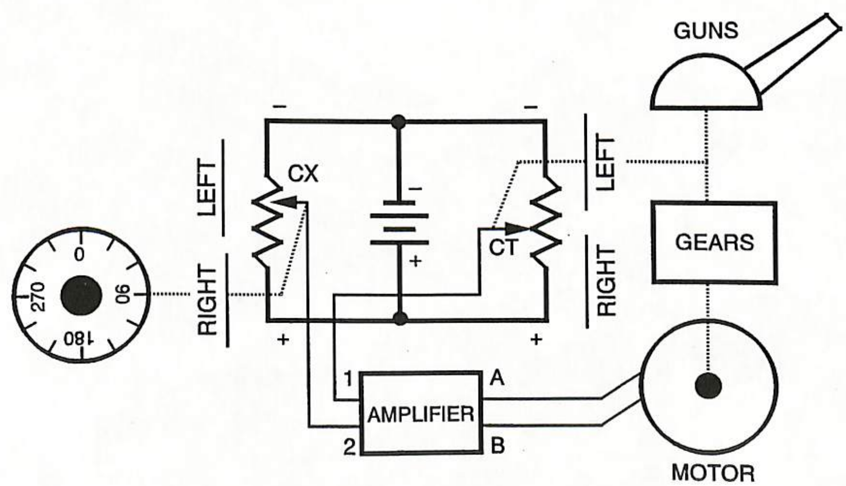


Figure 28. Gun Control Servo System

Positioning. Ideally, the output shaft should instantly and accurately match the motion and position of the input shaft. When an actual input signal is compared with a feedback response, a time lag results between the input signal and actual load movement that becomes larger as the weight of the load increases. This time lag can be decreased by increasing the gain of the servo amplifier. Increased gain makes the load move faster, which can cause it to pass (overshoot) the commanded position. When the system attempts to drive the load back in the other direction it could overshoot again.

This process also is called hunting. If the gain is set too high, the servo output tends to oscillate and is unstable, so gain is limited by stability requirements. A system with this problem must be stabilized to minimize or eliminate hunting. This is accomplished through damping. A compromise must be reached between the increased gain and the damping to have a smooth, efficiently operating system.

Damping. Damping can be obtained either by introducing a voltage in opposition to the signal voltage or by placing a physical restraint on the servo output. The actual function of an antihunting network is to reduce the amplitude and duration of oscillations that exist in the system. Every system has one or more natural oscillating frequencies, depending on load weight, designed speed, and other characteristics.

The simplest form of damping is friction damping, the application of friction to the output shaft or load proportional to the output velocity. The application of friction absorbs power from the motor which is dissipated in the form of heat. A pure friction damper absorbs excessive power from the system. A system having some characteristics of a friction damper, but less power loss, is used in actual practice. One such damper uses a friction clutch to couple a weighted flywheel to the output drive shaft of the servo motor. As the servo motor rotates, the clutch couples some of this motion to the flywheel. The flywheel overcomes inertia and gains speed, approaching the motor speed; it absorbs energy from the servo motor. The amount of energy stored in the flywheel is determined by its speed (velocity). Because of inertia, the flywheel resists any attempt to change its velocity.

As the correspondence (null) point of the system is approached, the error signal is reduced and the motor begins to slow down. As it attempts to keep the output shaft turning at the same speed, the flywheel releases some of its energy into the shaft, causing the first overshoot to be large. When the servo system drives past the point of correspondence, a new error signal is developed. The new error signal, of opposite polarity, causes the servo system motor to drive in the opposite direction. Once again the flywheel resists the motor movement and absorbs energy from the system. This causes a large reduction in the second overshoot and all subsequent overshoots of the system. The overall effect is to dampen oscillations about the point of correspondence and reduce the synchronizing time.

Another damper is the magnetic clutch, similar in function to the friction-clutch damper. The main difference between the two is the method used to couple the flywheel to the shaft of the servo motor.

There are two distinct types of magnetic-clutch dampers. The first uses a magnetic field to draw two friction-clutch plates together to produce damping in an action similar to friction-clutch action. The second version of the magnetic clutch uses a magnetic field generated by two sets of coils, or by one set of coils and the induced eddy currents resulting from rotation of the single set of coils near a conduction surface (the flywheel). Coupling in this clutch is made by the interaction of two magnetic fields. The two-coil or eddy-current type of magnetic clutch offers smoother operation than a pure friction clutch and has no problem of wear due to friction.

Time lag. A smooth, efficiently-operating servo system can be achieved only through compromise. Increasing the gain of the amplifier to reduce time lag has the drawback of increasing hunting. Overcoming this difficulty using friction damping solved the problem of hunting and smoothed out servo operation, but increased the servo load. This caused a large first overshoot and increased the time lag. Some form of damping useable with high amplification to obtain servo operation with minimum time lag is needed.

Error-rate damping. Error-rate damping anticipates the amount of overshoot. This form of damping corrects the overshoot by introducing a voltage in the error detector proportional to the rate of change of error signal. This voltage is combined with the error signal in the proper ratio to obtain the desired servo operation with reduced overshooting and minimum time lag. Error-rate voltages are generated by electromechanical devices or electrical networks in the equipment.

The advantages of error-rate damping are:

1. Maximum damping occurs when a maximum rate of change of error signal is present, usually as the servo load reverses direction.

2. Since a change in the signal causes damping, there is a minimum amount of damping either when no signal or a signal of constant strength is present.

SUMMARY

Synchro systems are used in everything from radio-controlled toys to nuclear reactors. They are in widespread use in military equipment. Basic synchro schematics resemble a three-phase wye generator with one rotor and three stator winding connections. Differential synchro schematics have three stator and three rotor winding connections. Connecting two synchros together forms a complete synchro system.

Synchros are used in automatic control systems for low-power, remotely-controlled shaft repositioning/indicating applications. They use an input on a knob to position a load or use an input from a load to position a dial or other indicator.

The output of a synchro system is not strong enough to drive a heavy load so a servo system is connected to the synchro output. The servo amplifier and servo motor are used to position the load. Damping increases the stability of the system.

13.3 Variable Frequency Drive Videos

- What is a VFD? (Variable Frequency Drive)
- Variable Frequency Drives Explained - VFD Basics IGBT inverter

13.4 Motor Manufacturer Specifications

13.4.1 RS-775126000E7 Motor

PG27 RS775-125 Motor with Encoder
The test motor was purchased here: www.andymark.com (PG27 RS775-125 Motor with Encoder)

Product Overview

This motor is the same motor included in our PG27 Gearmotor, and is being sold with the intention of replacing the motor without having to purchase the gearbox as well. This motor comes with the pinion gear and encoder attached and also includes the 2 mounting screws.

The back shaft of this motor features our Hall Effect Encoder, which can be removed if your application needs the long back shaft.

This motor, without the encoder, is am-2766, and is similar to am-2194, (same motor slightly shorter shaft). Both of these base motors are called out in the *FIRST* Robotics Competition manual as being legal for use in 2022.

Encoder Pinout:
- Pin 1: 5V DC
- Pin 2: Ground
- Pin 3: Channel A Output
- Pin 4: Channel B Output

Specifications

- Back Shaft Diameter: 0.2 in.
- Back Shaft Length: 0.45 in.
- No Load RPM: 5700 RPM
- Pulses Per Revolution: 7
- Shaft Diameter: 0.125 in.
- Stall Current: 15-20 AMPs
- Stall Torque: 35 oz-in
- Teeth: 11
- Voltage: 12 Volts DC
- Wattage: 44
- Weight: 0.86 lbs.

Motor dimensions pdf can be downloaded here: andymark.com (layout prints PG27 RS775-125 Motor with Encoder)

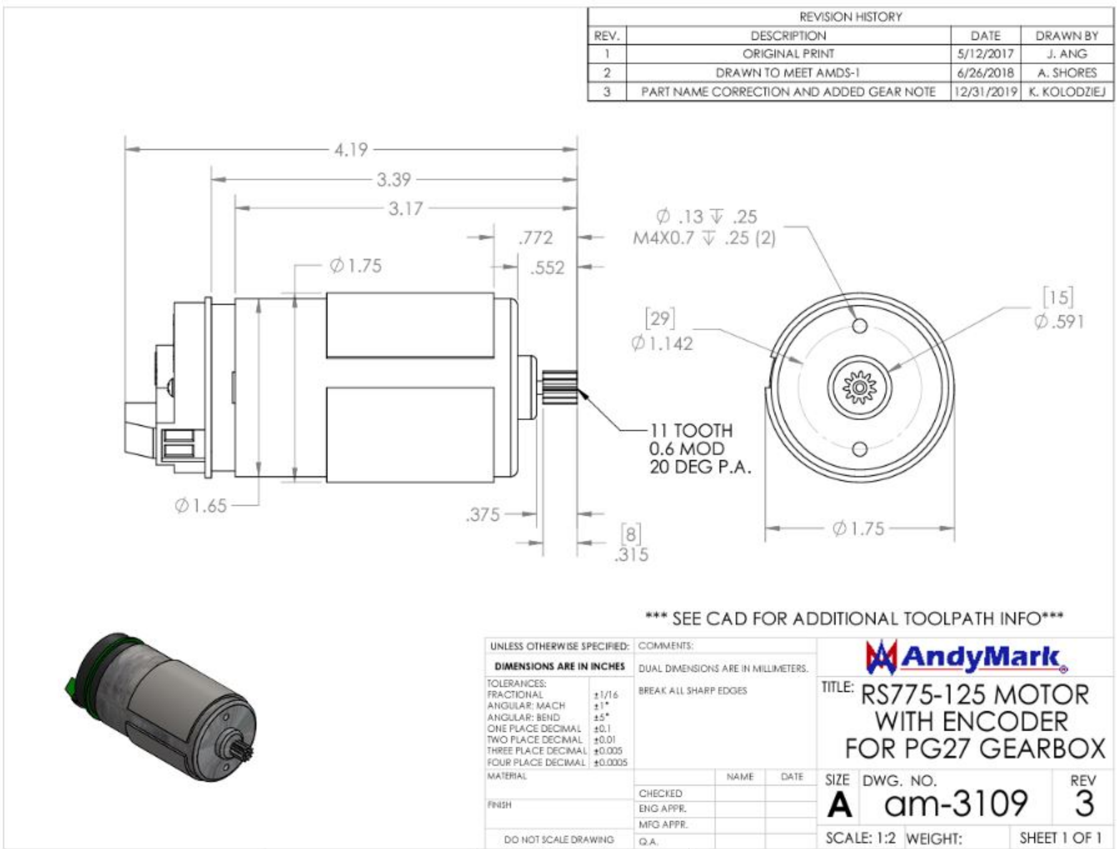


Figure 13.1: RS-775126000E7 Dimensions

13.4.2 TS-25GA370H-20 Motor

Encoder Metal Gearmotor 12V DC High Speed 300RPM Gear Motor with Encoder for Arduino and 3D Printers

The test motor was purchased here: www.amazon.com (Encoder Metal Gearmotor 12V DC High Speed 300RPM Gear Motor with Encoder for Arduino and 3D Printers)

Product type: DC Gear motor with two-channel Hall effect encoder

- Rated Voltage: 12V
- No-Load Speed: 300RPM
- No-Load Current: $\leq$ 0.15A
- Rated Torque: 0.5kg.cm
- Single Output 240 Pulses Per Revolution
- Gear Reduction Ratio: 1/20
- Each Loop Output Pulses: 12PPR 20×12=240PPR

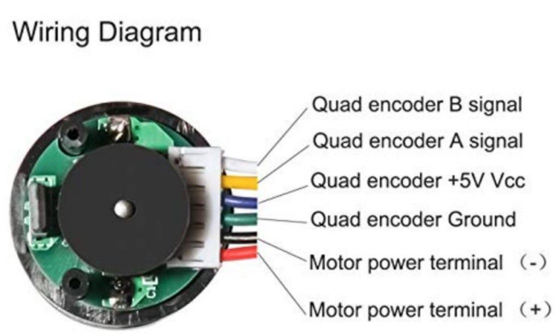


Figure 13.2: TS-25GA370H-20 Wiring

Motor Dimensions:

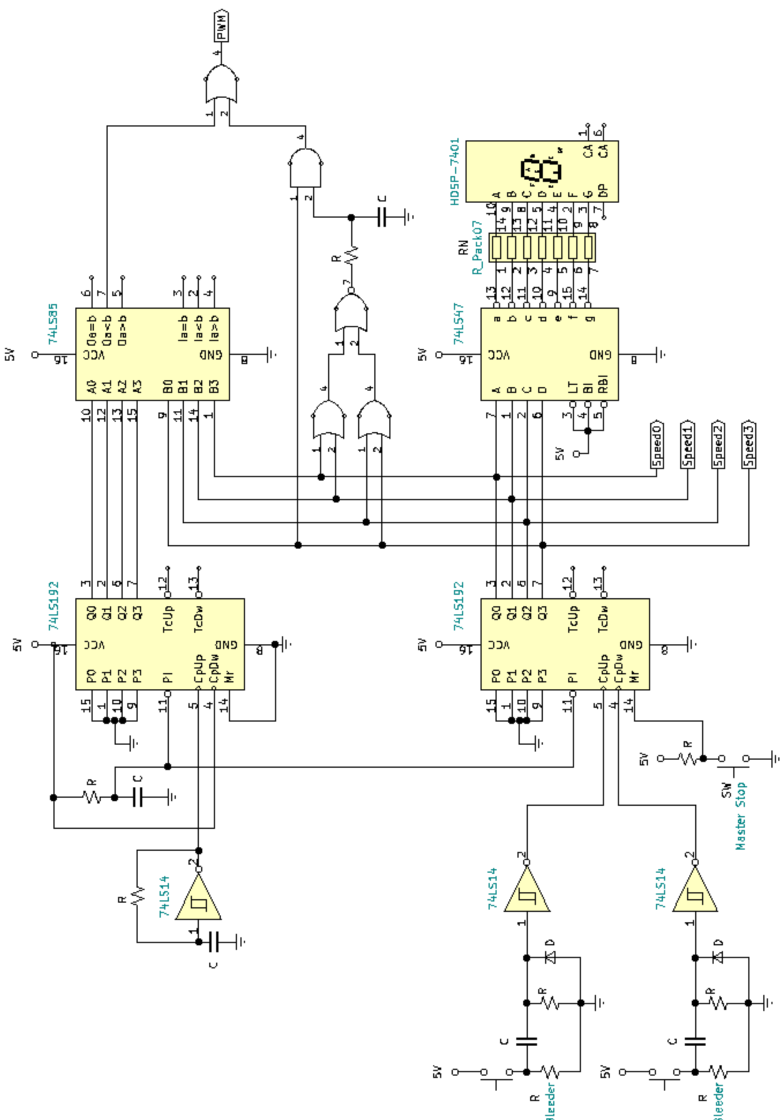
- Main Body: 66 x 25mm / 2.6in x 0.99in (L*D)
- Out Shaft: 11 x 4mm / 0.433 x 0.1575in (L*D) with 10 x 0.5mm / 0.39 * 0.017inches flat cut off

Figure 13.3: TS-25GA370H-20 Dimensions

13.5 H-Bridge Circuit Analysis

13.5.1 DC Motor PWM H-Bridge Schematic

RCET0251 MOTORS, PWM & H-BRIDGE TIM LEISHMAN



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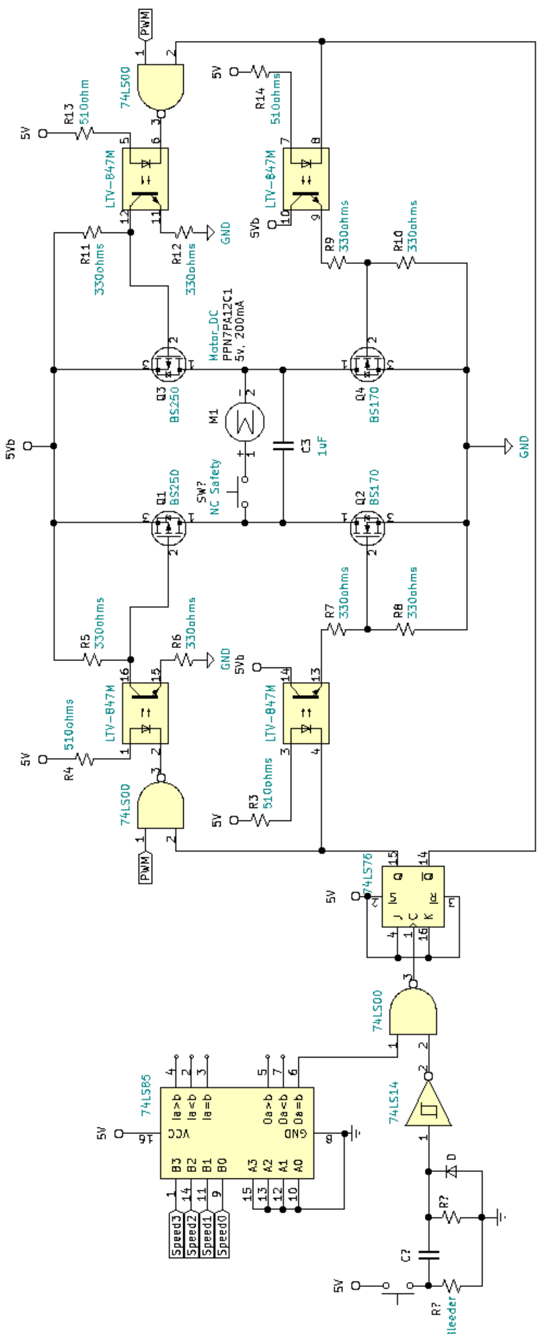
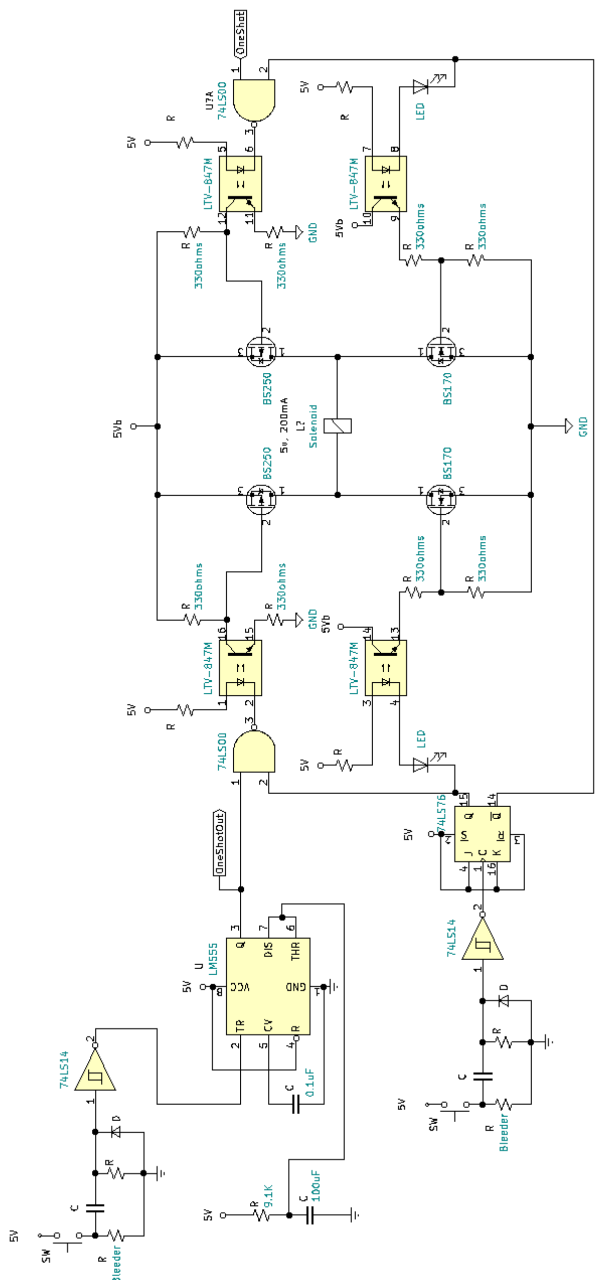

PWM
74LS00
R13
510ohm
5V
LTV-847M
R14
510ohms
R11
330ohms
R12
330ohms
GND
5Vb
R9
330ohms
R10
330ohms
Motor_DC
PPN7PA12C1
5v, 200mA
Q3
BS250
M1
Q4
BS170
C3
1uF
SW?
NC Safety
Q1
BS250
Q2
BS170
R5
330ohms
R6
330ohms
R7
330ohms
R8
330ohms
R4
510ohms
R3
510ohms
74LS76
74LS85
Speed3
Speed2
Speed1
Speed0
74LS14
C?
R?
Bleeder
D

13.5.2 Solenoid H-Bridge Schematic

RCET0251 MOTORS, H-BRIDGE SOLINOID TIM LEISHMAN



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13.6 Servo Parallax Documentation

PARALLAX

Web Store: www.parallax.com
Tutorials: learn.parallax.com
Sales: sales@parallax.com
Tech Support: support@parallax.com
Sales: (888) 512-1024
Educator Hotline: (916) 701-8625
Office: (916) 624-8333
Fax: (916) 624-8003

Parallax Continuous Rotation Servo (#900-00008)

The Parallax Continuous Rotation Servo provides bi-direcitonal continuous rotation with a linear response to pulse width. Two are used to power the drive wheels in the following Parallax robotics kits:

- BASIC Stamp Boe-Bot (#28132, #28832, #81031)
- SumoBot, Original (#27400) and WX (#32134)
- Shield-Bot with Arduino (#32335 & 81033)
- cyber:bot with micro:bit (#32700)

This document supports the Digital version of the Parallax Continuous Rotation Servo, released in summer 2022. It is designed to be a drop-in replacement of the previous analog version for the purpose of Parallax robotics; applications like adding this device to a digital bus are not supported. For the analog Parallax Continuous Rotation Servo previously sold, see the product guide version 2.2 available from the 900-00008 product page at www.parallax.com.

For best results, use two of the same type (analog or digital) when selecting or replacing servos for your robot.

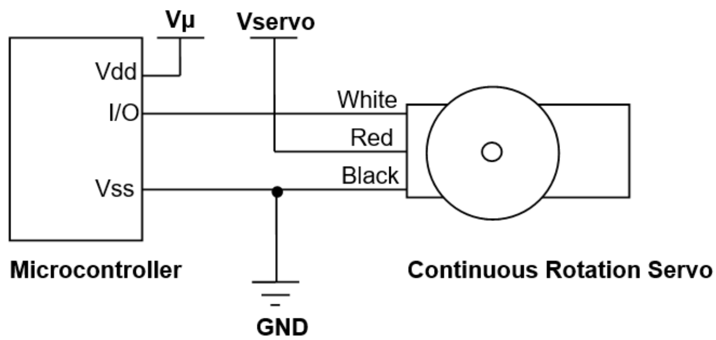
Features

- Bidirectional continuous rotation
- 0 to +/-50 RPM with 5 V supply
- Linear response to pulse-width for easy speed ramping
- Color coded cable (red VDD, white signal, black ground) with 3-pin 0.1" socket
- Accepts four mounting screws
- Easy to interface with any Parallax microcontroller or PWM-capable device

Specifications

- Gear material: POM (polyacetal)
- Spline: 3F 25-tooth
- Weight: 1.50 oz (42.5 g)
- Torque: 38 oz-in @ 6 V
- Power requirements: 4 to 6 VDC; Maximum current draw 140 +/- 50 mA at 6 VDC when operating in no load conditions, 15 mA when in static state
- Communication: pulse-width modulation
- Center pulse width: 1520 µs (before manual centering)
- Zero speed seadband: 12 µs
- Dimensions: approx 2.2 x 0.8 x 1.6 in (5.58x 1.9 x 4.06 cm) excluding servo horn
- Operating temperature range: 14 to 122 °F (-10 to +50 °C)

Electrical Connections



Vµ = microcontroller voltage supply

Vservo = 4 to 6 VDC, regulated or battery

I/O = PWM TTL or CMOS output signal, 3.3 to 5 V; < Vservo + 0.2 V

To use this servo with a Parallax robot, refer to the robot's documentation on learn.parallax.com. To use this servo with other devices, keep the following power precautions and other tips below in mind.

Power Precautions

- Power the servo with regulated 4-6 VDC, with a supply that can provide up to 500 mA as load may spike when the servo starts, stops, or turns. Four alkaline AA batteries or five NiMH rechargeable batteries in series are also acceptable as a supply for these servos.
- Servo current draw can spike while under peak load; be sure your application's regulator is rated to supply adequate current for all servos used in combination.
- Do not use this servo with an unregulated wall-mount supply. Such power supplies may deliver variable voltage far above the stated voltage.
- Do not try to power the servo directly through a microcontroller I/O pin.

Other Tips

- Calibrate (center) the servo with your project's microcontroller before permanently installing it into your application.
- Calibration requires access to the port in the side of the servo's case. For periodic recalibration, you may wish to design your project so the port remains accessible after installation.
- When using two servos for a small robot's drive wheels, use two of the same type (analog or digital).
- When using two servos for a small robot's drive wheels with servo horns pointed outwards, keep in mind that one servo must rotate clockwise while the other rotates counterclockwise at the same rate in order for the robot to drive straight.

Device Information

The Parallax continuous rotation servo relies on pulse width modulation to control the rotation speed and direction of the servo shaft. Before using the servo in a project, it is important to calibrate the servo's center setting in order to define the pulse width value at which the servo holds still (see the section Calibration – "Center" the Servo).

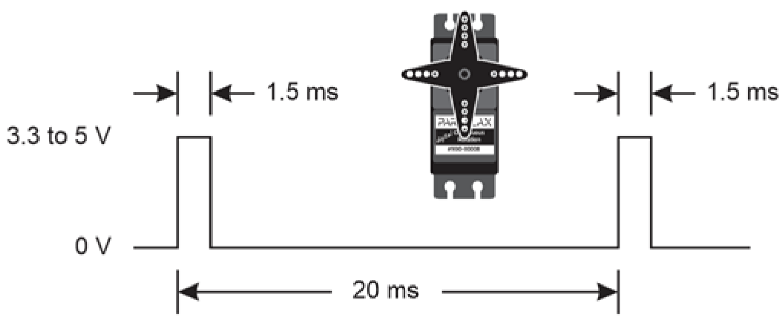
Specifications

Pin	Name	Description	Minimum	Typical	Maximum	Units
1 (White)	Signal	Input; TTL or CMOS	3.3	5.0	Vservo + 0.2	V
2 (Red)	Vservo	Power Supply	4.0	5.0	6.0	V
3 (Black)	Vss	Ground		0		V

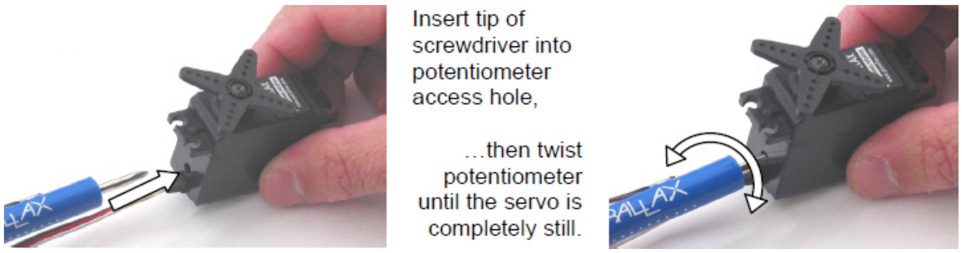
Calibration – "Center" the Servo

In most Parallax small robots, the servos are manually centered. Centering is a procedure that makes the servo stay still in response to a certain control signal.

Step 1: A microcontroller is programmed to send 1.5 ms (1500 µs) control pulses every 20 ms. In response to these pulses, a servo that has not been centered before typically responds by turning its output shaft (and horn) slowly clockwise.



Step 2: A #1 Phillips screwdriver is used to adjust the servo's trim potentiometer to make it stop turning. This establishes the 1.5 ms (1500 µs) as the signal to make the servo stay still.



NOTE: The above procedure does not ensure that the middle of the servo's deadband range is aligned with 1500 µs. For example, instead of staying still in response to pulses in the 1494 to 1506 µs range, it might instead stay still in response pulses in the 1490 µs to 1502 µs range. To more precisely center the servo around 1500 µs pulses, a microcontroller program can be written that repeatedly sends:

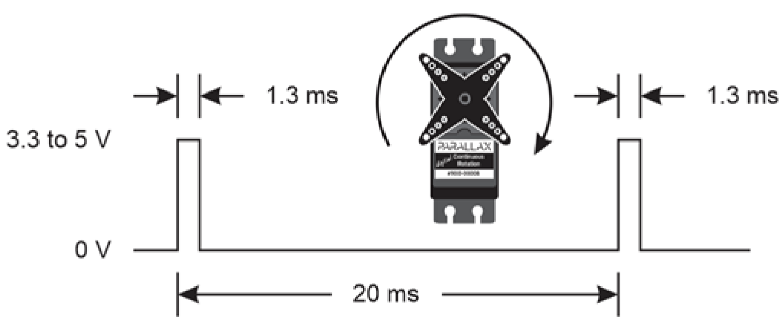
- 1485 µs pulses every 20 ms for 2 seconds
- 1515 µs pulses every 20 ms for 2 seconds.

Adjust the potentiometer with the screwdriver until the slow counterclockwise and clockwise velocities match.

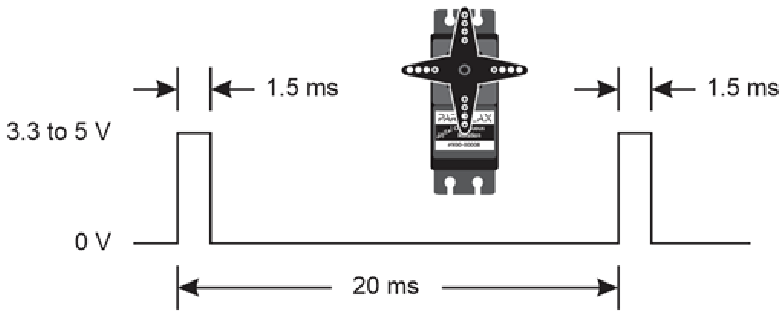
Communication Protocol

The Parallax Continuous Rotation Servo is controlled through pulse width modulation. Rotational speed and direction are determined by the duration of a high pulse, in the 1.3–-1.7 ms range. To rotate smoothly, the servo needs the control pulses to be repeated every 20 ms. Below are example timing diagrams for signals to make the servo turn full speed in both directions as well as stay still (assuming the centering procedure above has been completed):

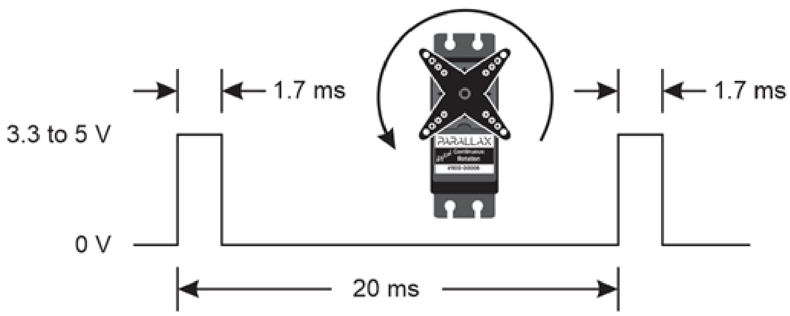
Full speed clockwise:



No rotation:

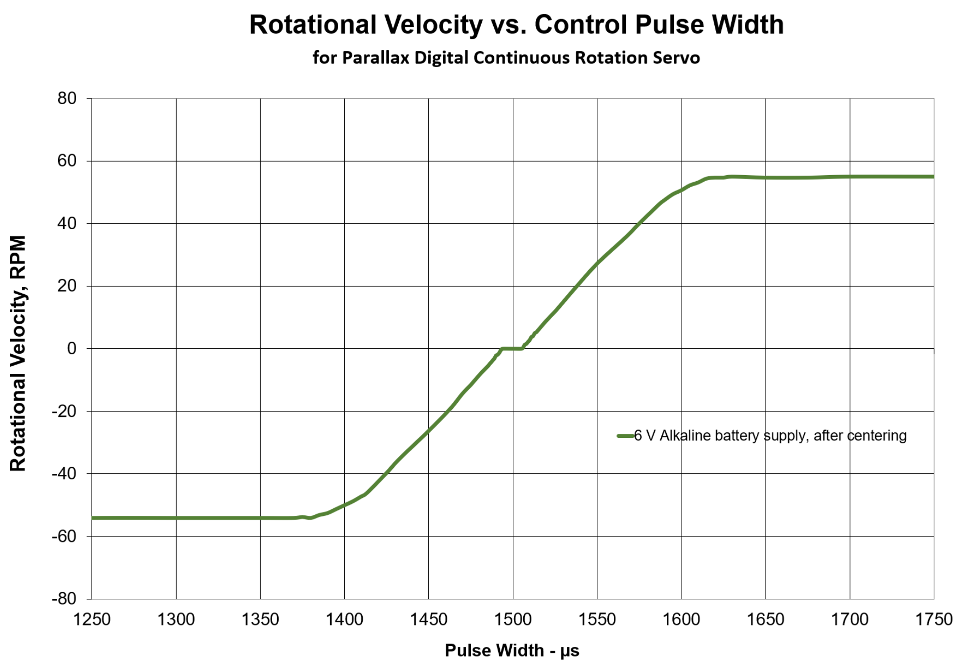


Full speed counterclockwise:



Control Servo Speed with Pulse Widths from its Transfer Function

The transfer function graph below shows the servo's velocity response to pulse width. Positive RPM values represent counterclockwise rotation; negative values represent clockwise rotation. With the servo precisely centered at 1500 µs, counterclockwise rotation speed increases linearly as the control pulse width is swept from 1506 to 1586 µs. Pulse widths at or above 1620 µs result in full speed. Likewise, clockwise rotation speed increases linearly with control pulse widths from 1494 to 1914 µs, and pulses with durations less than 1380 µs result in full speed.

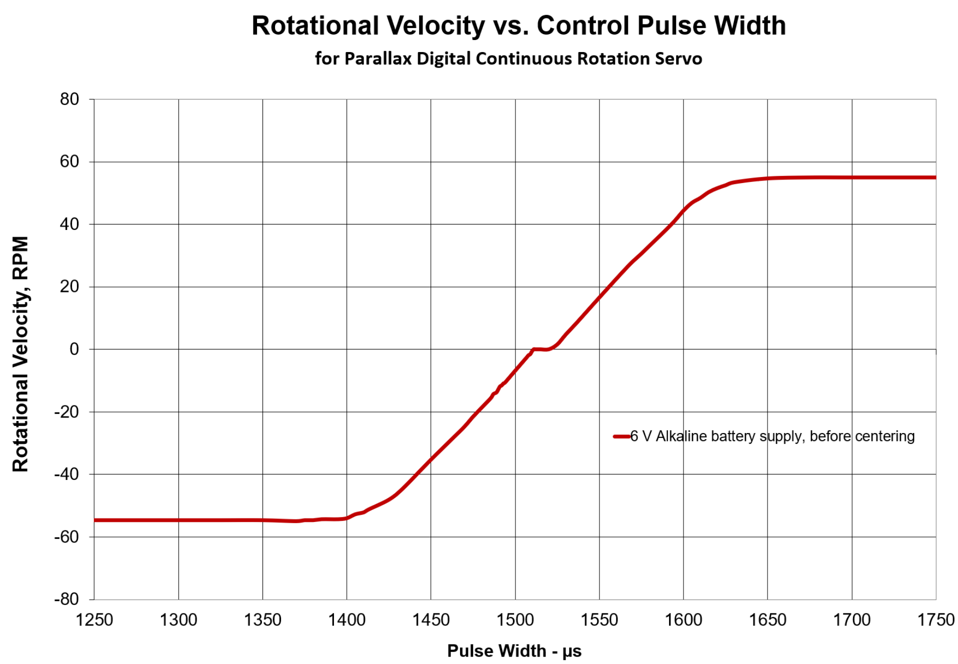


NOTES:

Variations in manufacturing, supply, temperature, and centering will affect your servo's actual values. Particularly, battery pack decay will affect the velocity response. The servos will run faster with new batteries, but the speed will decay as the batteries drain. With these variations in mind, supply voltage and control signal adjustment will be needed to suit your continuous rotation servo application.

Maximum RPM will vary with input voltage; 50 RPM @ 5 V is typical.

Before manual centering, the transfer function for a Parallax Digital Continuous Rotation Servo resembles this plot. The zero speed deadband is typically 12 µs wide, and centered at 1520 µs.



Resources and Downloads

Check for the latest servo documentation and resources on the #900-00008 product page at www.parallax.com. For servo use in a specific Parallax robot see the tutorials at learn.parallax.com.

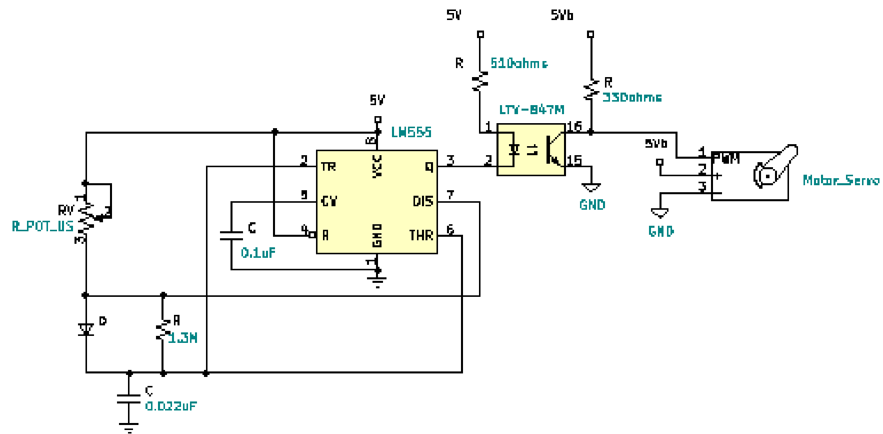
Revision History

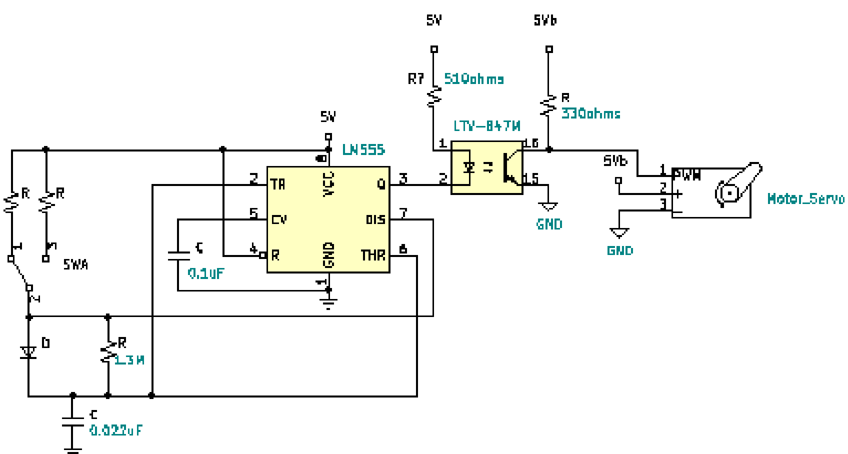
3.0 supports the digital version of the Parallax Continuous Rotation Servo, released in June 2022. See 2.2 for information on the previous analog version of this servo; it is available from the 900-00008 product page at www.parallax.com.

13.7 Servo PWM Circuit Analysis

13.7.1 Servo PWM Schematics

RCET0251 MOTORS, PWM SERVO TIM LEISHMAN





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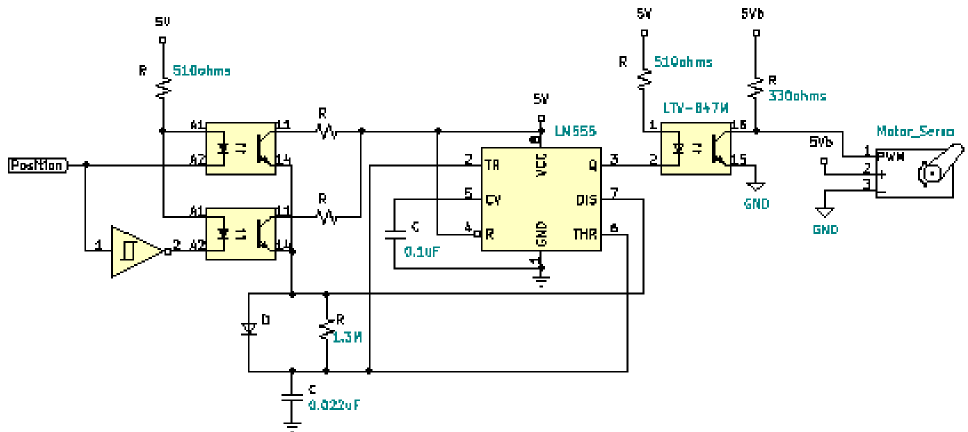
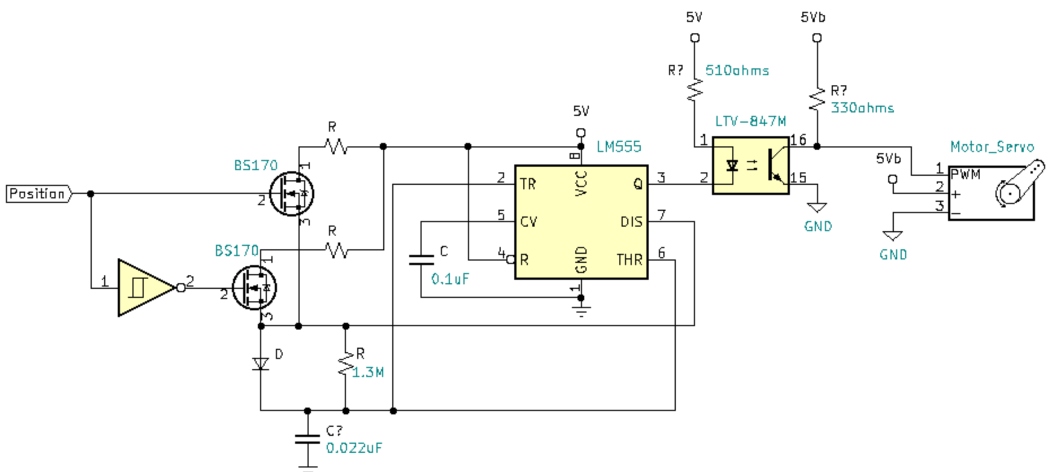

5V
R
510ohms
Position
R
R
5V
LM555
TR
VCC
Q
CV
DIS
R
GND
THR
C
0.1uF
D
R
1.3M
C
0.022uF
5V
R
510ohms
5Vb
R
330ohms
LTV-847M
GND
Motor_Servo
PWM
5Vb
GND


Position
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0.022uF
5V
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510ohms
5Vb
R?
330ohms
LTV-847M
GND
Motor_Servo
PWM
5Vb
GND

Week 14

Stepper Motors

14.1 Modern Control Technology - Kilian [1]

CHAPTER

8 Stepper Motors

OBJECTIVES

After studying this chapter, you should be able to:

- Explain what a stepper motor is, how it is different from a "regular" motor, and the applications it is used in.
- Understand the basic parts and operation of the three kinds of stepper motors: permanent magnet, variable reluctance, and hybrid.
- Differentiate between two-phase, three-phase, and four-phase stepper motors.
- Understand the different operational modes—single-step versus slew, single- and dual-phase excitation, half-step, and microstepping.
- Calculate the final position of a stepper motor, given the sequence of drive pulses.
- Explain the operation of stepper motor driver circuits.

INTRODUCTION

A **stepper motor** is a unique type of DC motor that rotates in fixed steps of a certain number of degrees. Step size can range from 0.9 to 90°. Figure 8.1 illustrates a basic stepper motor, which consists of a **rotor** and **stator**. In this case, the rotor is a permanent magnet, and the stator is made up of electromagnets (field poles). The rotor will move (or step) to align itself with an energized field magnet. If the field magnets are energized one after the other around the circle, the rotor can be made to move in a complete circle.

Stepper motors are particularly useful in control applications because the controller can know the exact position of the motor shaft without the need of position sensors. This is done by simply counting the number of steps taken from a known reference position. Step size is determined by the number of rotor and stator poles, and there is no **cumulative error** (the angle error does not increase, regardless of the number of steps taken). In fact, most stepper motor systems operate open-loop—that is, the controller sends the motor a determined number of step commands and assumes the motor goes to the right place. A common example is the positioning of the read/write head in a disk drive.

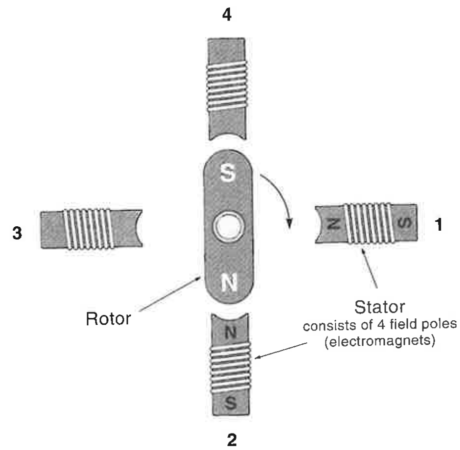


Figure 8.1
A PM 90° stepper motor.

Steppers have inherently low velocity and therefore are frequently used without gear reductions. A typical unit driven at 500 pulses/second rotates at only 150 rpm. Stepper motors can easily be controlled to turn at 1 rpm or less with complete accuracy.

There are three types of stepper motors: permanent magnet, variable reluctance, and hybrid. All types perform the same basic function, but some differences among them may be important in some applications.

8.1 PERMANENT MAGNET STEPPER MOTORS

The **permanent magnet (PM) stepper motor** uses a permanent magnet for the rotor. Figure 8.1 shows a simple PM stepper motor. The field consists of four poles (electromagnets). The motor works in the following manner: Assume the rotor is in the position shown with the south end up. When field coil 1 is energized, the south end of the rotor is attracted to coil 1 and moves toward it. Then field coil 1 is deenergized, and coil 2 is energized. The rotor pulls itself into alignment with coil 2. Thus, the rotor turns in 90° steps for each successive excitation of the field coils. The motor can be made to reverse by inverting the sequence.

One desirable property of the PM stepper motor is that the rotor will tend to align up with a field pole even when no power is applied because the PM rotor will be attracted to the closest iron pole. You can feel this "magnetic tug" if you rotate the motor by hand; it is called the detent torque, or residual torque. The detent torque is a desirable property in many applications because it tends to hold the motor in the last position it was stepped to, even when all power is removed.

As mentioned earlier, one big advantage of the stepper motor is that it can be used open-loop—that is, by keeping track of the number of steps taken from a known point, the exact shaft position is always known. Example 8.1 demonstrates this.

♦ EXAMPLE 8.1

A 15°/step stepper motor is given 64 steps CW (clockwise) and 12 steps CCW (counterclockwise). Assuming it started at 0°, find the final position.

Solution

After completing 64 steps CW and 12 steps CCW, the motor has ended up 52 steps CW ($64 - 12 = 52$). Because there are 24 15°-steps per revolution ($360°/15° = 24$),

$$\frac{52}{24} = 2\frac{1}{6} = 2 \text{ rev} + \frac{1}{6} \text{ rev}$$

$$= 2 \text{ rev} + \frac{360°}{6}$$

$$= 2 \text{ rev} + 60°$$

Therefore, the motor has made two complete revolutions and is now sitting at 60° CW from where it started. ♦

Effect of Load on Stepper Motors

For the open-loop concept to work, the motor must actually step once each time it's commanded to. If the load is too great, the motor may not have enough torque to make the step. In such a case, the rotor would probably rotate a little when the step pulse was applied but then fall back to its original position. This is called **stalling**. If feedback is not used, the controller has no way of knowing a step was missed.

Within each step, the torque developed by the stepper motor is dependent on the shaft angle. In fact, the torque on the rotor is actually zero when it is exactly aligned with an energized field coil. Figure 8.2 illustrates how the motor can only provide torque when the rotor is *not* aligned. The first frame of Figure 8.2 shows a rotor pole approaching an energized field pole. The actual force of attraction is between the south (S) end of the rotor and the north (N) end of the field pole. As the rotor pole approaches the field pole, the attraction force (*F*) gets stronger but the torque component (*T*) gets weaker. When the rotor is pointing directly at the field pole (last frame in Figure 8.2), the torque component is zero. In practice, this means that the rotor may come to a stop before it is completely aligned with the energized field pole, at the point where the diminishing step torque just equals the load torque.

For the simple motor under discussion (Figure 8.1), the maximum torque occurs when the rotor is 90° (one step) away from the field pole (first frame of Figure 8.2.) It might seem that we should just plan to let the rotor lag one step behind the energized field pole to take advantage of the maximum torque, but this approach might cause the

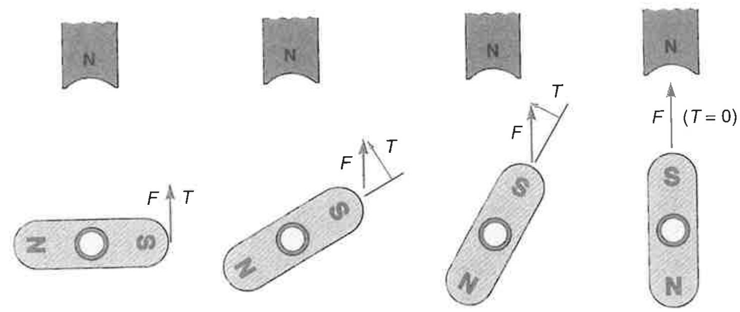


Figure 8.2
Torque goes to zero as the rotor aligns with the field pole.

motor to take a step backward instead of forward. Consider the situation in Figure 8.3(a) where the rotor has been stepping CCW and we have allowed it to lag a full step behind the energized pole (currently, pole 1). The next pole to be energized in the CCW sequence is pole 2 [Figure 8.3(b)]. The first problem here is that the rotor is pointing directly away from pole 2, so there will be little or no torque exerted. The second problem is that, in this balanced condition, the rotor will be equally attracted in either direction and we cannot reliably predict if it will turn CW or CCW.

For proper operation, *the rotor lag must not be to allowed exceed one-half the step size*, which would be 45° for the motor illustrated in Figure 8.3. This solves the preceding problems—namely, the motor will always turn in the direction it's supposed to, and it

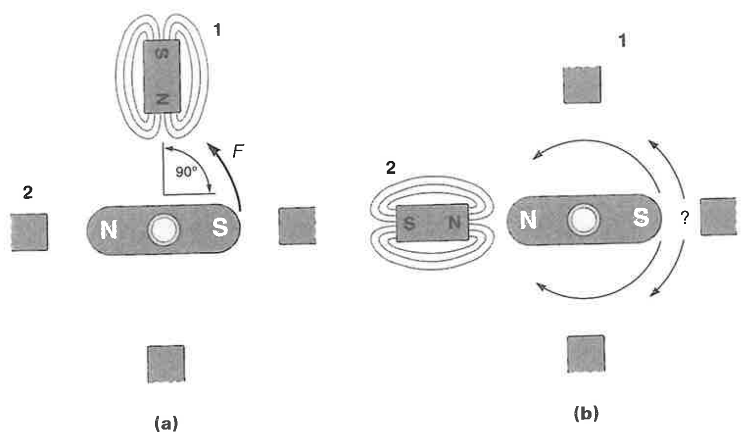


Figure 8.3
Illustrating what would happen if the rotor was allowed to lag a full step behind the field poles.

will not stall. (Recall that stalling occurs when the motor is too weak to take a step.) In practical terms, the dynamic torque, which is the power available when the motor is running, may only be about half of the maximum *static torque* (the torque required to displace the rotor when stopped). There is an exception to this rule: When the rotor is stepping rapidly (called *slewing*), the inertia can be counted on to keep the rotor going in the right direction. Slewing is discussed in the next section.

Modes of Operation

The stepper motor has two modes of operation: single step and slew. In the **single-step mode** or bidirectional mode, the frequency of the steps is slow enough to allow the rotor to (almost) come to a stop between steps. Figure 8.4 shows a graph of position versus time for single-step operation. For each step, the motor advances a certain angle and then stops. If the motor is only lightly loaded, overshoot and oscillations may occur at the end of each step as shown in the figure.

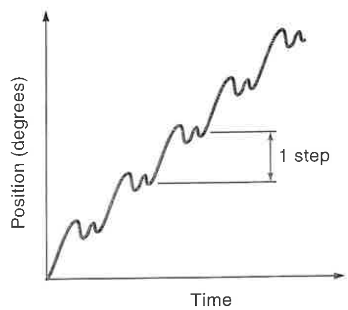


Figure 8.4
Position vs time for single-step mode.

The big advantage of single-step operation is that each step is completely independent from every other step—that is, the motor can come to a dead stop or even reverse direction at any time. Therefore, the controller has complete and instantaneous control of the motor's operation. Also, there is a high certainty that the controller will not lose count (and hence motor position) because each step is so well defined. The disadvantage of single-step mode is that the motion is slow and "choppy." A typical single-step rate is 5 steps/second which translates to 12.5 rpm for a 15°/step motor.

In the **slew mode**, or unidirectional mode, the frequency of the steps is high enough that the rotor does not have time to come to a stop. This mode approximates the operation of a regular electric motor—that is, the rotor is always experiencing a torque and rotates in a smoother, continuous fashion. Figure 8.5 shows a graph of position versus time for the slew mode. Although the individual steps can still be discerned, the motion is much less choppy than in single-step mode.

A stepper motor in the slew mode cannot stop or reverse direction instantaneously. If attempted, the rotational inertia of the motor would most likely carry the rotor ahead

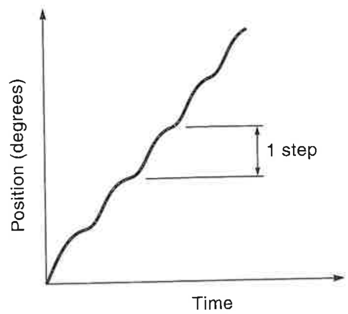


Figure 8.5
Position vs time for the slew mode.

a few steps before it came to rest. The step-count integrity would be lost. It is possible to maintain the step count in the slew mode by slowly ramping up the velocity from the single-step mode and then ramping down at the end of the slew. This means the controller must know ahead of time how far the motor must go. Typically, the slew mode is used to get the motor position in the "ballpark," and then the fine adjustments can be made with single steps. Slewing moves the motor faster but increases the chances of losing the step count.

Figure 8.6 shows the torque–speed curves for both the single-step and slew modes. The first observation is that available load torque diminishes as the stepping rate rises (this is true of all DC motors). Also, for the single-step mode, the price paid for the ability to stop or reverse instantaneously is less torque and speed. Looking along the x-axis, notice three different kinds of torques. The **detent torque** is the torque required to overcome the force of the permanent magnets (when the power is off). It is the little tugs

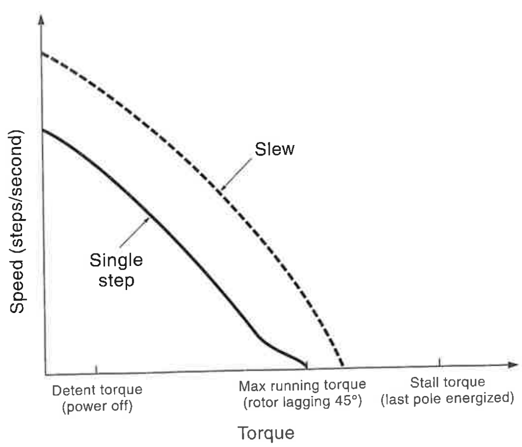


Figure 8.6
Torque–speed curves for single-step and slew modes.

you feel if you manually rotate the unpowered motor. The **dynamic torque**, which is the maximum running torque, is obtained when the rotor is lagging behind the field poles by half a step. The highest stall torque shown in Figure 8.6 is called the **holding torque** and results when the motor is completely stopped but with the last pole still energized. This is really a detent type of torque because it represents the amount of external torque needed to rotate the motor "against its wishes."

♦ EXAMPLE 8.2

A stepper motor has the following properties:

Holding torque: 50 in. · oz
Dynamic torque: 30 in. · oz
Detent torque: 5 in. · oz

The stepper motor will be used to rotate a 1-in. diameter printer platen (Figure 8.7). The force required to pull the paper through the printer is estimated to not exceed 40 oz. The static weight of the paper on the platen (when the printer is off) is 12 oz. Will this stepper motor do the job?

Solution

The torque required to rotate the platen during printing can be calculated as follows:

$$\text{Force} \times \text{radius} = 40\text{ oz} \times 0.5\text{ in.} = 20\text{ in.}\cdot\text{oz}$$

Therefore, the motor, with 30 in. · oz of dynamic torque, will be strong enough to advance the paper.

The torque on the platen from just the weight of the paper is calculated as follows:

$$\text{Force} \times \text{radius} = 12\text{ oz} \times 0.5\text{ in.} = 6\text{ in}\cdot\text{oz}$$

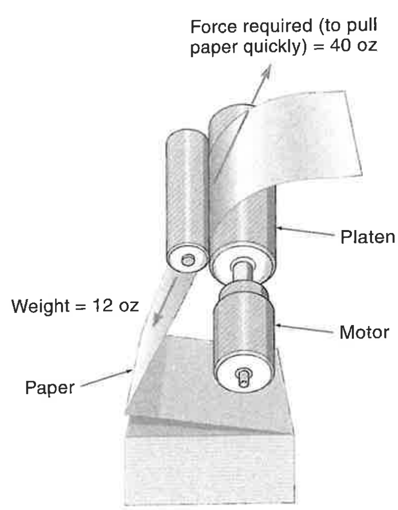


Figure 8.7
A stepper motor driving a printer platen (Example 8.2).

When the printer is on, the powered holding torque of 50 in. · oz is more than enough to support the paper. However, when the printer is off, the weight of the paper exceeds the detent torque of 5 in. · oz, and the platen (and motor) would spin backward. Therefore, we conclude that this motor is not acceptable for the job (unless some provision such as a ratchet or brake is used to prevent back spinning). ◆

Excitation Modes for PM Stepper Motors

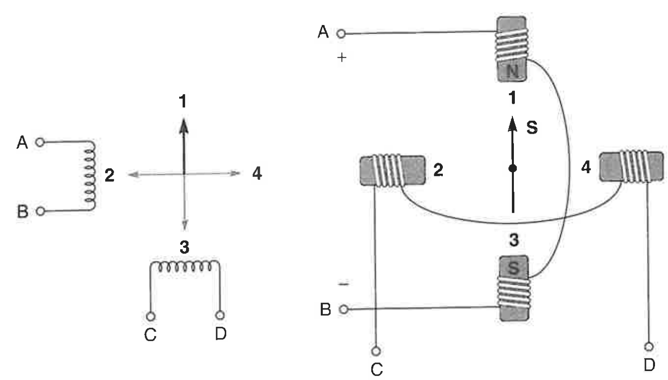
Stepper motors come with a variety of winding and rotor combinations. In addition, there are different ways to sequence energy to the field coils. All these factors determine the size of each step. **Phase** refers to the number of separate winding circuits. There are two-, three-, and four-phase steppers, which are discussed next.

Two-Phase (Bipolar) Stepper Motors

The **two-phase** (bipolar) **stepper motor** has only two circuits but actually consists of four field poles. Figure 8.8(a) shows the motor symbol, and Figure 8.8(b) shows how it is wired internally. In Figure 8.8(b), circuit AB consists of two opposing poles such that when voltage is applied (+A −B), the top pole will present a north end to the rotor and the bottom pole will present a south end. The rotor would tend to align itself vertically (position 1) with its south pole up (because, of course, opposite magnetic poles attract).

The simplest way to step this motor is to alternately energize either AB or CD in such a way as to pull the rotor from pole to pole. If the rotor is to turn CCW from position 1, then circuit CD must be energized with polarity C+ D−. This would pull the rotor to position 2. Next, circuit AB is energized again, but this time the polarity is reversed (−A +B), causing the bottom pole to present a north end to the rotor, thereby pulling

Figure 8.8
A two-phase (bipolar) stepper motor.



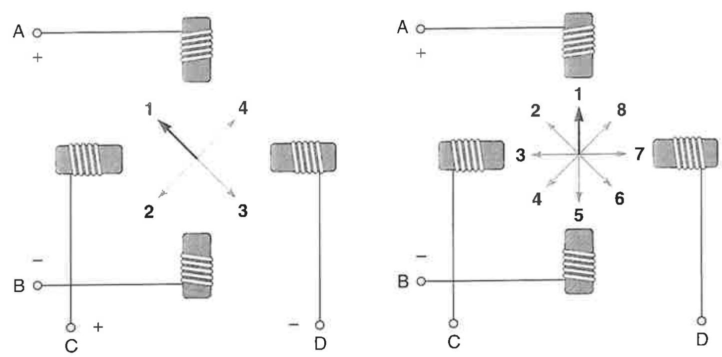
(a) Symbol **(b)** Wiring diagram

it to position 3. The term **bipolar** applies to this motor because the current is sometimes reversed. The voltage sequence needed to rotate the motor one full turn is shown below. Reading from top to bottom gives the sequence for turning CCW, reading from bottom up gives the CW sequence:

Circuit	Position
A+ B−	1
C+ D−	2
A− B+	3
C− D+	4

Another way to operate the two-phase stepper is to energize both circuits at the same time. In this mode, the rotor is attracted to two adjacent poles and assumes a position in between. Figure 8.9(a) shows the four possible rotor positions. The excitation sequence for stepping in this dual mode is as follows:

Circuits			Position
A+ B−	and	C+ D−	1
A− B+	and	C+ D−	2
A− B+	and	C− D+	3
A+ B−	and	C− D+	4



(a) Dual excitation (energized for position 1)

(b) 8-step drive using "half steps"

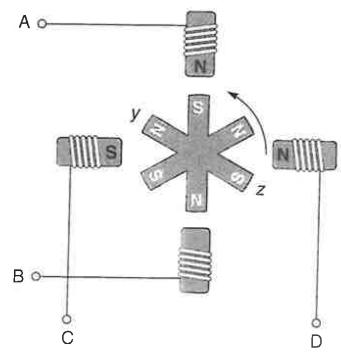
Figure 8.9
Additional operating modes for stepper motors.

Having two circuits on at the same time produces considerably more torque than the single-excitation mode; however, more current is consumed, and the controller is more complex.

Both methods produce **four-step drives**, that is, four steps per cycle. By alternating the single- and dual-excitation modes, the motor can be directed to take **half-steps**, as shown in Figure 8.9(b). Positions 1, 3, 5, and 7 are from the single-excitation mode, and positions 2, 4, 6, and 8 are from the dual-excitation mode. When driven this way, the motor takes eight steps per revolution and is called an **eight-step drive**. This is desirable for some applications because it allows the motor to have twice the position resolution. Even smaller steps are possible with a process called *microstepping*, which is discussed later in the chapter.

The motor described thus far in this section steps 90° in the four-step mode (and 45° in the eight-step mode). PM stepper motors commonly have a smaller step, as low as 30°. This is done by increasing the poles in the rotor. Figure 8.10 shows a 30° stepper motor; the rotor has six rotor poles. Assume that field poles AB have been energized, pulling the rotor into the position shown. Next, circuit CD is energized. Rotor poles *yz* will be attracted to poles CD and will have to rotate only 30° to come into alignment.

Figure 8.10
A 30° stepper motor with a six-pole rotor.



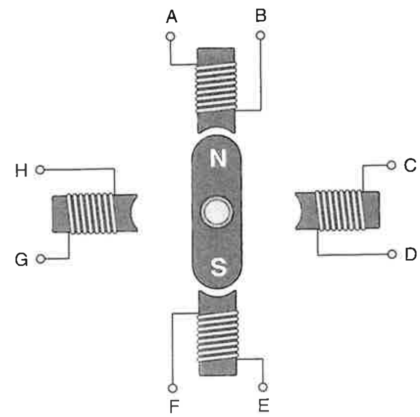
Four-Phase (Unipolar) Stepper Motors

The **four-phase** (unipolar) is the most common type of stepper motor (Figure 8.11). The term *four-phase* is used because the motor has four field coils that can be energized independently, and the term **unipolar** is applied because the current always travels in the same direction through the coils. The simplest way to operate the four-phase stepper motor is to energize one phase at a time in sequence (known as *wave drive*). To rotate CW, the following sequence is used:

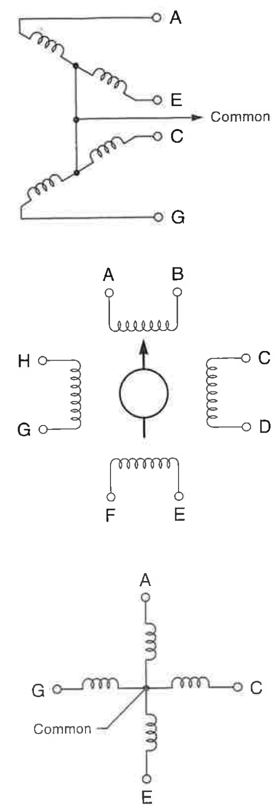
A B
C D
E F
G H

Compared with the two-phase bipolar motor, the four-phase motor has the advantage of simplicity. The control circuit of the four-phase motor simply switches the poles on and off in sequence; it does not have to reverse the polarity of the field coils. (However, the two-phase motor produces more torque because it is pushing and pulling at the same time.)

Figure 8.11
A four-phase (unipolar) stepper motor.



(a) Wiring diagram

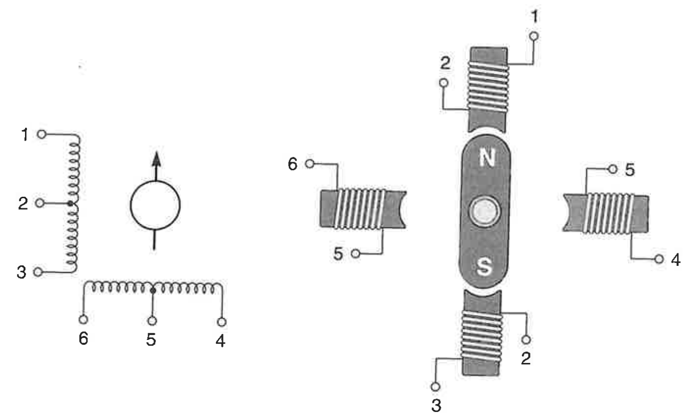


(b) Symbols

The torque of a four-phase stepper motor can be increased if two adjacent coils are energized at the same time, causing the rotor to align itself between the field poles [similar to that shown in Figure 8.9(a)]. Although twice the input energy is required, the motor torque is increased by about 40%, and the response rate is increased. By alternating single- and dual-excitation modes, the motor steps in half-steps, as shown in Figure 8.9(b).

Constructing motors so they can be used in either a two- or four-phase mode is common practice. This is done by bringing out two additional wires (from the two-phase motor) that are internally connected to points between the opposing field coils. Figure 8.12(a) shows the symbol for this type of motor, and Figure 8.12(b) shows the interior motor wiring. When such a motor is used in the two-phase mode, the center taps (terminals 2 and 5) are not used. When the motor is operated in the four-phase mode, the center taps become a *common return*, and power is applied to terminals 1, 4, 3, and 6 as required.

Figure 8.12
A four-phase stepper with center tap windings.

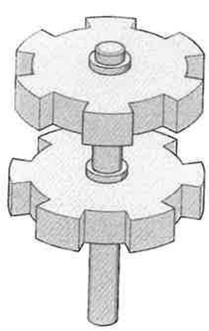


(a) Symbols (b) Wiring diagram

Available PM Stepper Motors

Almost all PM stepper motors available today have smaller step sizes than the simplified motors discussed so far. These motors are made by stacking two multipolled rotors (offset by one-half step), as illustrated in Figure 8.13. Typical step sizes for four-phase PM stepper motors are 30°, 15°, and 7.5°. Figure 8.14 is a specification sheet for a family of PM stepper motors. For example, the PF35–48C (continuous duty) motor has 48 steps/revolution (which is 7.5°), requires 133 mA at 12 V, and has a holding torque of 2.78 in. · oz.

Figure 8.13
Stacked-rotor design allows smaller steps.



8.2 VARIABLE RELUCTANCE STEPPER MOTORS

The **variable reluctance (VR) stepper motor** does not use a magnet for the rotor; instead, it uses a toothed iron wheel [see Figure 8.15(b)]. The advantage of not requiring the rotor to be magnetized is that it can be made in any shape. Being iron, each rotor tooth is

PF35 Series		Models					
		PF35-48				PF35-24	
Excitation Mode		2-2					
Step Angle	(°)	7.5				15	
Step Angle Tolerance	(%)	±5					
Steps per Revolution		48				24	
Rating		Continuous			Intermittent	Continuous	
Letter Designator		C	D	Q	C	C	D
Winding Type		Unipolar	Unipolar	Bipolar	Unipolar	Unipolar	Unipolar
DC Operating Voltage	(V)	12	5	5	24	12	5
Operating Current	(mA/ø)	133	313	310	266	133	313
Winding Resistance	(Ω/ø)	90	16	17	90	90	16
Winding Inductance	(mH/ø)	48	8.9	12	48	48	8.9
Holding Torque	(oz-in)	2.78		3.25	3.88	2.08	
Rotor Inertia	(oz-in²)	24.1×10^{-3}					
Starting Pulse Rate, Max.	(pps)	500		400	680	310	
Slewing Pulse Rate, Max.	(pps)	530		500	770	410	
Ambient Temp Range, Operating	(°C)	-10~+50					
Temperature Rise	(°C)	55			-	55	
Weight	(oz)	2.8					

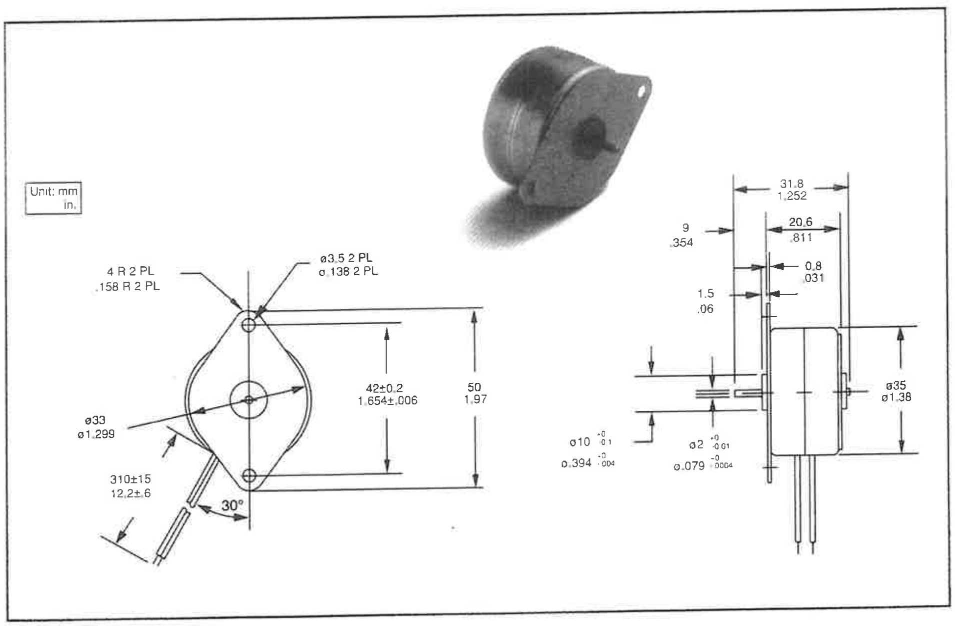
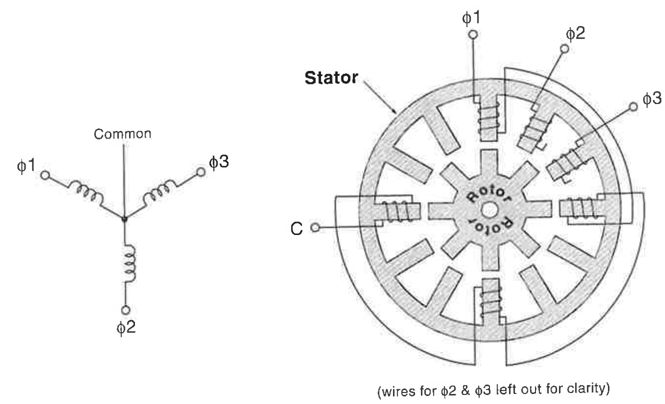


Figure 8.14 PM stepper motors. (Courtesy of Kollmorgen Motion Technologies Group)

Figure 8.15
A three-phase VR stepper motor (15° step). (Wires for Ø2 and Ø3 left out for clarity.)



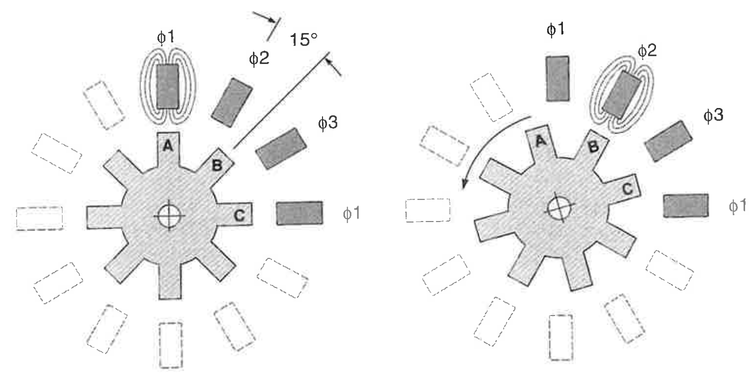
(a) Symbol **(b)** Construction

attracted to the closest energized field pole in the stator, but not with the same force as in the PM motor. This gives the VR motor less torque than the PM motor.

A VR motor usually has three or four phases. Figure 8.15(a) shows a typical **three-phase stepper motor**. The stator has three field pole circuits: Ø1, Ø2, and Ø3. Figure 8.15(b) shows that the actual motor has 12 field poles, where each circuit energizes four windings; you can see this by closely observing the Ø1 wire in Figure 8.15(b). Notice that the rotor has only 8 teeth even though there are 12 teeth in the stator. Therefore, the rotor teeth can never line up "one for one" with the stator teeth, a fact that plays an important part in the motor's operation.

Figure 8.16 illustrates the operation of the VR stepper motor. When circuit Ø1 is energized, the rotor will move to the position shown in Figure 8.16(a)—that is, a rotor tooth (A) is lined up with the Ø1 field pole. Next, circuit Ø2 is energized. Rotor tooth B,

Figure 8.16
A 15° three-phase VR stepper motor (only four field poles shown for clarity.)



(a) Pole ϕ1 energized **(b)** Pole ϕ2 energized being the closest, is drawn toward it [Figure 8.16(b)]. Notice that the rotor has to move only 15° for this alignment. If circuit Ø3 is energized next, the rotor would continue CCW another 15° by pulling tooth C into alignment.

The step angle of the VR motor is the difference between the rotor and stator angles. For the motor of Figure 8.16, the angle between the field poles is 30°, and the angle between the rotor poles is 45°. Therefore, the step is 15° (45° − 30° = 15°). By using this design, the VR stepper motor can achieve very small steps (less than 1°). Small step size is often considered to be an advantage because it allows for more precise positioning.

The VR stepper motor has a number of functional differences when compared with the PM type. Because the rotor is not magnetized, it is weaker than a similar-sized PM stepper motor. Also, it has no detent torque when the power is off, which can be an advantage or disadvantage depending on the application. Finally, because of the small step size and reduced detent torque, the VR stepper motor has more of a tendency to overshoot and skip a step. This is a serious matter if the motor is being operated open-loop, where position is maintained by keeping track of the number of steps taken. To solve the problem, some sort of damping may be required. This can be done mechanically by adding friction or electrically by providing a slight braking torque with adjacent field poles.

8.3 HYBRID STEPPER MOTORS

The **hybrid stepper motor** combines the features of the PM and VR stepper motors and is the type in most common use today. The rotor is toothed, which allows for very small step angles (typically 1.8°), and it has a permanent magnet providing a small detent torque even when the power is off.

Recall that the step size of a PM motor is limited by the difficulty in making a multipole magnetized rotor. There is simply a limit to the number of different magnetizations that can be imposed on a single iron rotor. The VR stepper motor gets around this by substituting iron teeth (of which there can be many) for magnetized poles on the rotor. This approach allows for a small step angle, but it sacrifices the strength and detent torque qualities of the PM motor. The hybrid motor can effectively magnetize a multitoothed rotor and thus has the desirable properties of both the PM and VR motors.

Figure 8.17 illustrates the internal workings of the hybrid motor, which is considerably more complicated than the simple PM motor. The rotor consists of two toothed wheels with a magnet in between—one wheel being completely north in magnetization and the other being completely south. For each step, two opposing teeth on the north wheel are attracted to two south field poles, and two opposing teeth on the south wheel are attracted to two north field poles. The internal wiring is more complicated than it is for the PM or VR motors, but to the outside world this motor is just as simple to control.

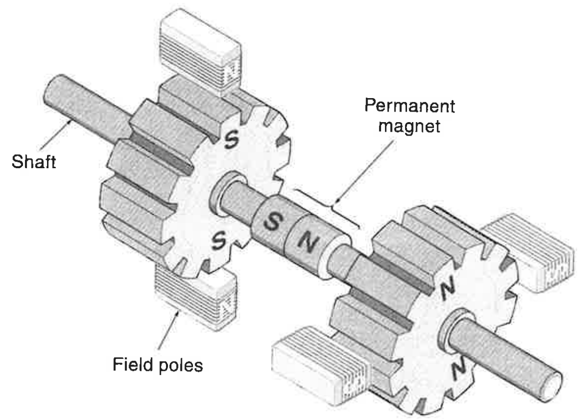


Figure 8.17
Internal construction of the hybrid stepper motor (only 2 poles per stator shown for clarity).

The theory of operation of the hybrid motor is similar to the VR motor in that the rotor and stator have a different number of teeth, and for each step, the closest energized teeth are pulled into alignment. However, the principles of magnetics require that, at any one time, half the poles be north and the other half be south. To maintain the magnetic balance, each pole must be able to switch polarity so that it can present the correct pole at the correct time. This is accomplished in one of two ways: For a bipolar motor, the applied voltage must be reversed by the driver circuit (similar to the two-phase PM motor). On the other hand, a unipolar motor has two separate windings of opposite direction on each field pole (called a *bifilar winding*), so each pole can be a north or a south. Therefore, the unipolar hybrid motor does not require a polarity-reversing driver circuit.

Figure 8.18 shows a specification sheet for a family of 1.8° hybrid stepping motors. For example, the 11-SHBD-45AB draws 0.3 A at 13.8 V and has a holding torque of 9.5 in. · oz, a running (dynamic) torque of 5.9 in. · oz, and a detent torque (unpowered) of 0.36 in. · oz. Unloaded, it can step at a rate of 1385 steps/minute.

8.4 STEPPER MOTOR CONTROL CIRCUITS

Figure 8.19 shows the block diagram for a stepper motor driving circuit. The *controller* decides on the number and direction of steps to be taken (based on the application). The *pulse sequence generator* translates the controller's requests into specific stepper motor coil voltages. The *driver amplifiers* boost the power of the coil drive signals. It should be

Bulletin SM 11/1.8

Size 11 1.8° stepper motors

Compact, light weight, high resolution stepper motors with high torque-to-size ratio

Features and Benefits

- Small size: 1.067" diameter
  1.500" long (excl. shaft)
- ±5% accuracy
- Half or full step (200 or 400 steps per revolution)
- High torque-to-size ratio
- Ball bearing construction
- Direct drive - no gearing
- Light weight
- Low cost

Typical applications

- Medical equipment
- Avionic instruments
- Robotics and automation
- Scanners
- Office equipment
- Battery operated equipment
- Chart recorders
- Test equipment
- Laser optics

STEPPER MOTOR SPECIFICATIONS	11-SHBD-45AB	11-SHBD-47AB	11-SHBD-49AB
INDEX ANGLE	1.8°±5%	1.8°±5%	1.8°±5%
FUNCTION	2ø HYBRID	4ø HYBRID	4ø HYBRID
STEPS PER REVOLUTION	200	200	200
INPUT VOLTAGE (DC)	13.8 REF.	9.6 REF.	14
INPUT CURRENT PER PHASE (AMPS) ± 10%	0.300	0.300	0.311
DC RESISTANCE PER PHASE (OHMS) ± 10%	46	32	45
INDUCTANCE PER PHASE (MH) REF.	51.3	22.4	28.4
NO LOAD RESPONSE RATE (PPS) MIN.	1140 (2)	875 (2)	810 (2)
NO LOAD SLEW RATE (PPS) MIN.	1385 (2)	1130 (2)	1225 (2)
HOLDING TORQUE (OZ-IN) MIN.	9.5 (2)	6.2 (2)	7.8 (2)
DYNAMIC TORQUE (OZ-IN) MIN.	5.9 (2)	3.9 (2)	4.3 (2)
DETENT TORQUE (OZ-IN) REF.	0.36	0.30	0.30
SHAFT RADIAL PLAY MAX.	0.0006 (0.015)	0.0006 (0.015)	0.0006 (0.015)
SHAFT END PLAY MAX. (1)	0.005 (0.13)	0.005 (0.13)	0.005 (0.13)
ROTOR INERTIA (GM-CM^2) REF.	3.3	3.8	3.6

(1) Shaft end play is spring loaded toward front of unit. (2) Measured with two phases on, L/R.

Figure 8.18 Hybrid stepper motors. (Courtesy of Litton Clifton Precision) clear that the stepper motor is particularly well suited for digital control; it requires no digital-to-analog conversion, and because the field poles are either on or off, efficient class C driver amplifiers can be used.

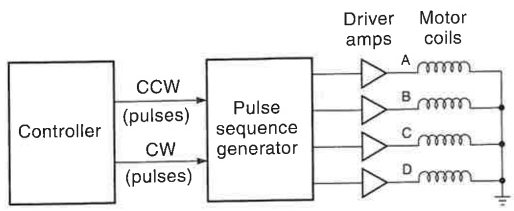


Figure 8.19
Block diagram of stepper motor control circuit.

Controlling the Two-Phase Stepper Motor

Controlling the two-phase bipolar stepper motor requires polarity reversals, making it more complicated than four-phase motor controllers. Figure 8.20 shows a two-phase stepper motor. The two circuits are designated AB and CD. The timing diagram shows the required waveforms for A, B, C, and D (CCW rotation). Looking down the position 1 column in Figure 8.20(b), we see A is positive and B is negative, so current will flow from A to B in circuit AB. Meanwhile C and D are both negative, effectively turning off circuit CD. For position 2 in the timing diagram, C is positive, and D is negative; causing current to flow from C to D in circuit CD while coil AB is completely off, and so on, for positions 3 and 4.

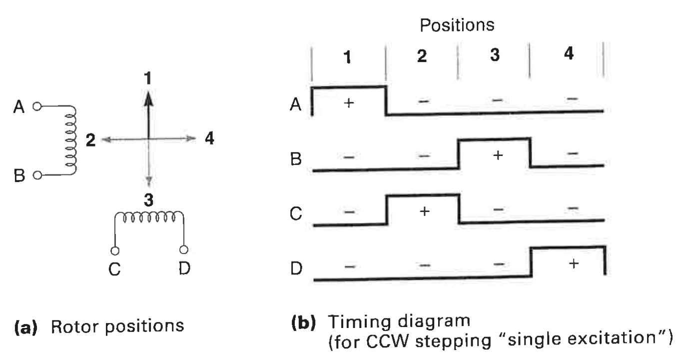


Figure 8.20
Two-phase (bipolar) stepper motor operation.

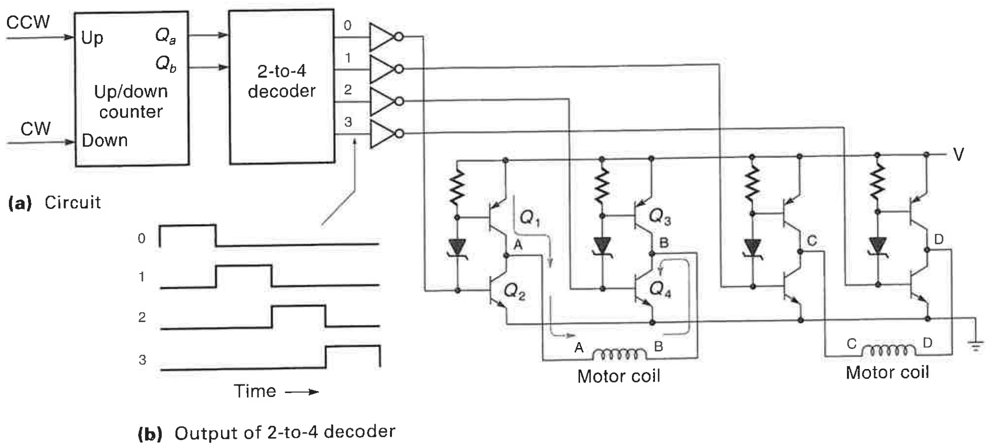


Figure 8.21 Complete interface circuit for a two-phase (bipolar) stepper motor.

A digital circuit such as the one shown in Figure 8.21(a) can be used to generate the timing waveforms. The up/down counter is a 2-bit counter that increments for each pulse received on its up input and decrements for each pulse received on the down input. Q_a and Q_b of the up/down counter are decoded in a 2-to-4 decoder. Because the counter is always in one of four states (00, 01, 10, 11), one (and only one) of the four decoder outputs is "high" at any particular time. Figure 8.21(b) shows the output of the decoder when the counter counts up (a result of CCW pulses from the controller).

The next task is to connect the timing signals from the decoder in such a way as to drive the motor coils. This can be accomplished with the power amplifier circuit shown on the right side of Figure 8.21(a). Notice there are four complementary–symmetry drivers, one for each end of each motor coil. When Q_1 and Q_4 are on, the current can flow through the motor in the direction shown. On the other hand, when Q_3 and Q_2 are on, the polarity is reversed, and the current flows the opposite direction through the motor. Finally, if Q_1 and Q_3 are off, no current flows through the motor coil.

The four outputs of the decoder (which must be inverted in this case) control the four complementary–symmetry transistor circuits. The resistor and diode in each circuit cause the upper transistor to be on when the lower transistor is off, and vice versa. Trace through the circuit for each step of the decoder, and you will see that the timing diagram of Figure 8.20(b) is reproduced. This arrangement will provide for the motor to step CCW when the counter counts up. When the counter counts down, the sequence will be backward, and the motor will step CW.

Controlling the Four-Phase Stepper Motor

The electronics needed to drive the four-phase stepper motor is simpler than for a two-phase motor because polarity reversals are not required. Figure 8.22 identifies the coils in a four-phase motor and shows the timing diagram for simple single-excitation

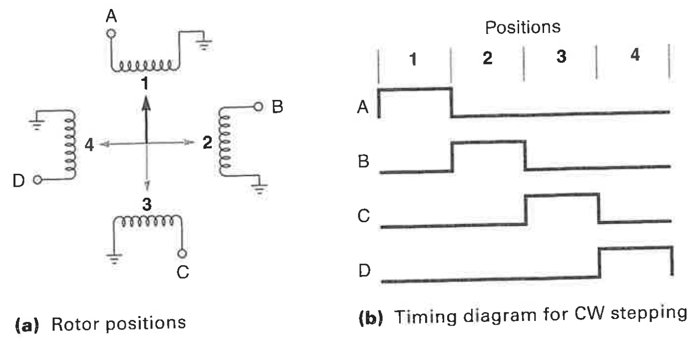


Figure 8.22
Four-phase (unipolar) stepper motor operation.

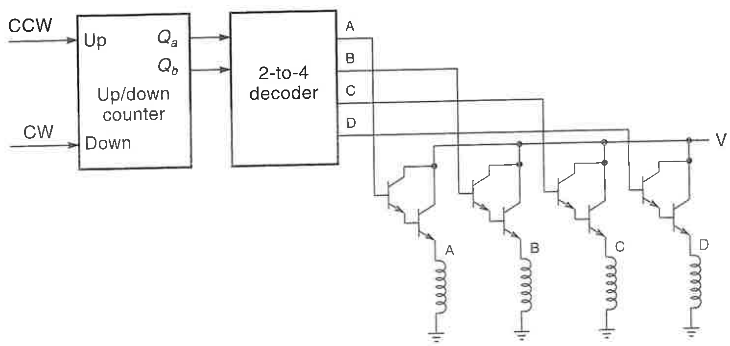


Figure 8.23
Complete interface for a four-phase stepper motor (simplified for clarity).

CW stepping. The timing is very straightforward and can easily be generated with the counter and decoder circuit shown in Figure 8.23.

The outputs of the decoder (Figure 8.23) are connected to four Darlington driver transistors. As timing signals A, B, C, and D go high in sequence, the corresponding transistors are turned on, energizing coils in the motor. The necessary four Darlingtons are available on a single IC such as the Allegro ULN-2064B (Figure 8.24). These amplifiers can supply up to 1.5 A and can be driven directly from TTL 5-V logic. (Note the diode to allow an escape path for current in the coil when the transistor is turned off.)

ICs designed specifically to drive stepper motors contain both the timing logic and power drivers in one package. One example is the Allegro UCN-5804B (Figure 8.25). The basic inputs are the step input (pin 11) and direction (pin 14). The motor will advance one step for each pulse applied to the step input pin, and the logic level on the direction pin determines if rotation will be CW or CCW. Notice that the output transistors are in the common emitter configuration (called *open collector*). The motor coils should be connected between the output pins and the supply voltage (as shown). When an output transistor turns on, it completes the circuit by providing a path to

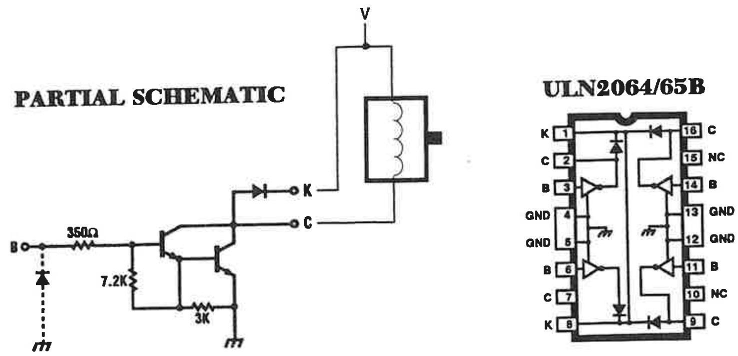
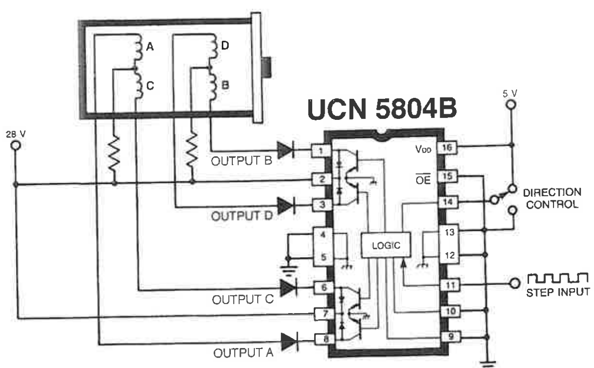


Figure 8.24 The Allegro ULN-2064B with four Darlington 1.5-A switches. (Courtesy of Allegro MicroSystems)

Figure 8.25 A unipolar stepper motor translator/driver (Allegro UCN-5804B). (Courtesy of Allegro MicroSystems)



(a) Driver circuit

WAVE-DRIVE SEQUENCE

← DIRECTION = L / → DIRECTION = H

Half Step = L, One Phase = H				
Step	A	B	C	D
POR	ON	OFF	OFF	OFF
1	ON	OFF	OFF	OFF
2	OFF	ON	OFF	OFF
3	OFF	OFF	ON	OFF
4	OFF	OFF	OFF	ON

TWO-PHASE DRIVE SEQUENCE

← DIRECTION = L / → DIRECTION = H

Half Step = L, One Phase = L				
Step	A	B	C	D
POR	ON	OFF	OFF	ON
1	ON	OFF	OFF	ON
2	ON	ON	OFF	OFF
3	OFF	ON	ON	OFF
4	OFF	OFF	ON	ON

HALF-STEP DRIVE SEQUENCE

← DIRECTION = L / → DIRECTION = H

Half Step = H, One Phase = L				
Step	A	B	C	D
POR	ON	OFF	OFF	OFF
1	ON	OFF	OFF	OFF
2	ON	ON	OFF	OFF
3	OFF	ON	OFF	OFF
4	OFF	ON	ON	OFF
5	OFF	OFF	ON	OFF
6	OFF	OFF	ON	ON
7	OFF	OFF	OFF	ON
8	ON	OFF	OFF	ON

(b) Modes of operation

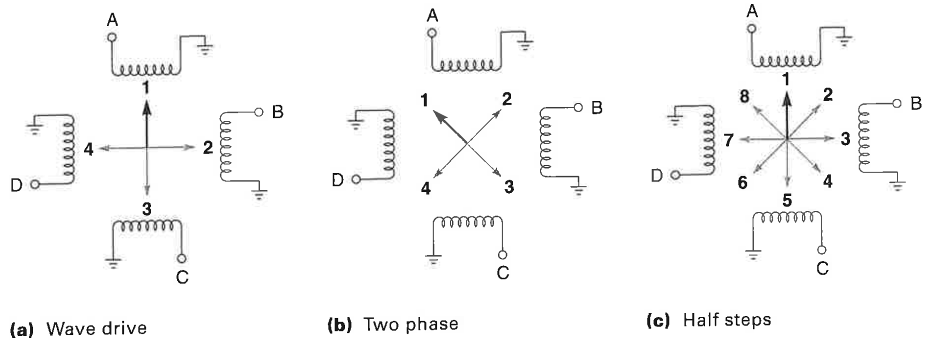


Figure 8.26 Three modes of operation for the Allegro UCN-5804B.

ground for the motor coil current. The three operating modes of the UCN-5804B are given in tabular form in Figure 8.25(b), and will be explained using Figure 8.26:

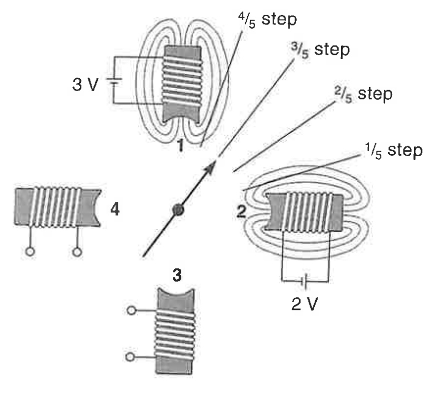
1. Figure 8.26(a) shows how the motor responds to the wave-drive sequence, a single-excitation mode where the coils A, B, C, and D are energized one at a time in sequence.
2. Figure 8.26(b) shows how the motor responds to the two-phase drive sequence, a dual-excitation mode where two adjacent phases are energized at the same time to give more torque.
3. Figure 8.26(c) shows how the motor responds to the half-step drive sequence, where the operation alternates between single- and dual-excitation modes, yielding eight half-steps per cycle.

Microstepping

Microstepping, a technique that allows a stepper motor to take fractional steps, works by having two adjacent field poles energized at the same time, similar to half-steps described earlier. In microstepping the adjacent poles are driven with different voltage levels, as demonstrated in Figure 8.27(a). In this case, pole 1 is supplied with 3 V and pole 2 with 2 V, which causes the rotor to be aligned as shown—that is, three-fifths of the way to pole 1. Figure 8.27(b) shows the voltages (for poles 1 and 2) to get five microsteps between each "regular" step. The different voltages could be synthesized with pulse-width modulation (PWM). The most commonly used microstep increments are 1/5, 1/10, 1/16, 1/32, 1/125, and 1/250 of a full step. Another benefit of microstepping (for delicate systems) is that it reduces the vibrational "shock" of taking a full step—that is, taking multiple microsteps creates a more "fluid" motion.

Two other points on microstepping: It does not require a special stepper motor, only special control circuitry, and the actual position of the rotor (in a microstepping system) is very dependent on the load torque.

Figure 8.27
Microstepping.



(a)

Pole 1	Pole 2	Position
5 V	0 V	Pole 1 (full step)
4 V	1 V	4/5 step
3 V	2 V	3/5 step
2 V	3 V	2/5 step
1 V	4 V	1/5 step
0 V	5 V	Pole 2 (full step)

(b)

Improving Torque at Higher Stepping Rates

It is important that the stepper motor develop enough torque with each step to drive the load. If it doesn't, the motor will stall (not step). When steps are missed, the controller no longer knows the exact position of the load, which may render the system useless.

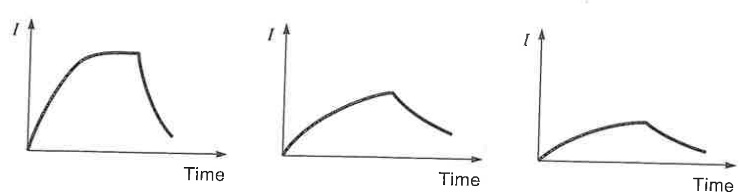
At higher stepping rates, two problems occur. First, if the load is accelerating, extra torque is needed to overcome inertia; second, the available motor torque actually diminishes at higher speeds. Recall that motor torque is directly proportional to motor current and that the average current decreases as the stepping rate increases. This is illustrated in Figure 8.28, which shows the motor current for three stepping rates. The problem is that the rate of change of current is limited by the circuit-time constant τ.

$$\tau = \frac{L}{R} \tag{8.1}$$

where

τ = time constant
L = motor inductance
R = motor coil resistance

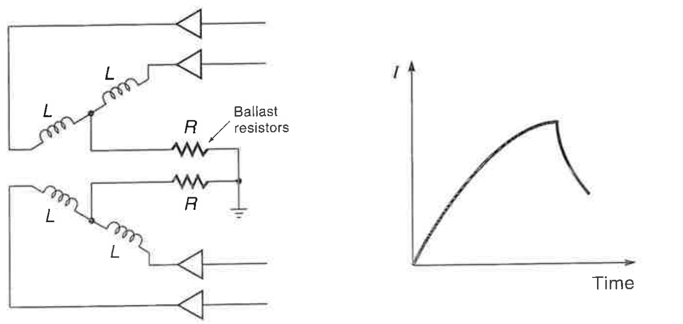
Figure 8.28
Coil current as a function of stepping speed.



(a) 100 steps/second (b) 500 steps/second (c) 1000 steps/second

You can see that as the stepping rate increases the current cannot build up in the field coils to as great a value. If we could reduce the value of the motor-time constant, the current could build up faster. One way to do this would be to increase the value of R in Equation 8.1. This can easily be done by adding external resistors (R) in series with the motor coils as shown in Figure 8.29(a). Such resistors are called **ballast resistors**, and their purpose is to improve the torque output at higher stepping rates (it also limits the current, which may be important in some cases). Stepper motor driver circuits that use ballast resistors are called **L/R drives**. Figure 8.29(b) shows the motor current with the ballast resistor added (for a rate of 1000 steps/second). Compare this with the last graph in Figure 8.28 to see the improvement.

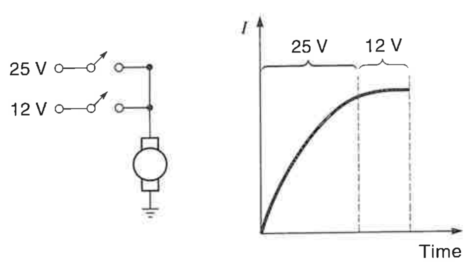
Figure 8.29
Effect of adding an external ballast resistor.



(a) Motor coils and ballast resistors **(b)** Coil current at 1000 steps/second with ballast resistor

Another way to improve motor torque at higher stepping rates is to use **bilevel drive**. In this approach, a high voltage is momentarily applied to the motor at the beginning of the step to force a fast in-rush of current. Then a lower voltage level is switched on to maintain that current. Figure 8.30(a) shows a simplified circuit to provide bilevel drive.

Figure 8.30
Principle of a bilevel drive.

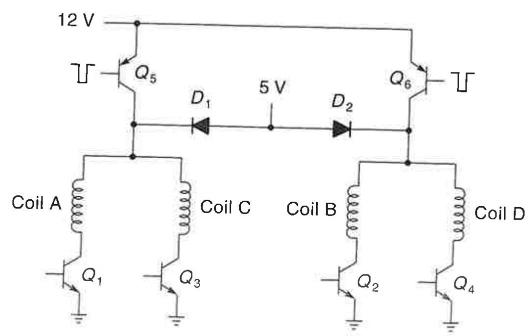


(a) **(b)**

The 25-V circuit is switched on, and the current rises rapidly [Figure 8.30(b)]. When the desired current level is reached, the 25-V circuit is switched off, and the 12-V circuit is switched on, which keeps the current at the desired level for the rest of the step time.

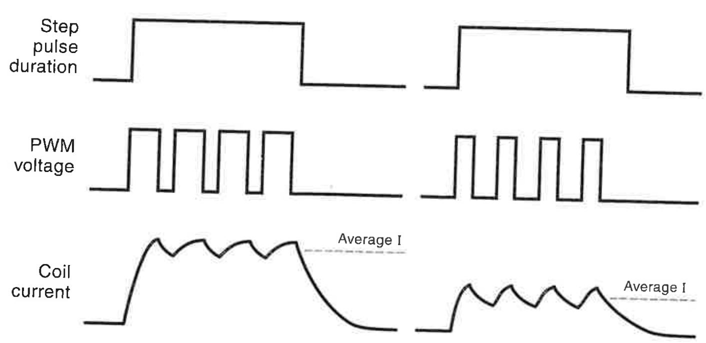
Figure 8.31 shows a bilevel-drive interface circuit. In this case, the higher voltage is 12 V, and the lower voltage is 5 V. The 12-V is switched through either Q_5 or Q_6 in response to a pulse from a timing circuit (not shown). The 5-V is applied through D_1 and D_2. These diodes keep the 12-V pulses from backing up into the 5-V power supply. The bilevel drive is more complex but allows the stepper motor to have more torque at higher stepping rates.

Figure 8.31
A bilevel-drive circuit.



Another approach for providing higher torque at faster stepping rates is the **constant current chopper drive**. Using PWM techniques, this driver circuit can deliver almost the same current to the motor at all speeds. A chopper-drive waveform is shown in Figure 8.32 and works in the following manner: A relatively high voltage is switched to the motor coil, and the current is monitored. When the current reaches a specified level, the voltage is cut off. After a short time, the voltage is reapplied, and the current again

Figure 8.32
A PWM (chopper drive) used to regulate stepper motor coil current.



(a) Higher average coil current **(b)** Lower average coil current increases, only to be cut off, and so on. Thus, in the same way that a thermostat can maintain a constant temperature by switching the furnace on and off, the chopper drive maintains a constant average current (within each drive pulse) by rapidly switching the voltage on and off. In summary, the chopper drive is another technique for providing good torque at high stepping rates. Stepper motor driver ICs are available, such as the Allegro A2919SB, with built-in PWM constant current capability.

8.5 STEPPER MOTOR APPLICATION: POSITIONING A DISK DRIVE HEAD

Example 8.3 illustrates many of the principles presented in this chapter and extends the discussion to show how software can control a stepper motor.

◆ EXAMPLE 8.3

A 30° four-phase stepper motor drives the read/write head on a floppy disk drive (Figure 8.33). The in-and-out linear motion is achieved with a leadscrew connected directly to the motor shaft. Each magnetic track on the disk is 0.025 in. apart (40 tracks/in.). The leadscrew has 20 threads/in.

The motor is driven by the UCN-5804B stepper motor interface IC. This IC requires only two inputs: step input and direction. A computer will supply these signals in response to toggle switch settings. A front panel contains eight toggle switches; seven are used to input (in binary) the number of tracks to move, and the eighth switch specifies direction—in or out. Write a program in BASIC that will cause the motor to step the number of tracks and direction specified by the switch settings.

Solution First we need to find the number of steps required to advance one track on the disk. If the leadscrew has 20 threads per inch, then rotating it one revolution (360°) will advance it 1 thread, which is 1/20 in. The following equation was set up by multiplying all the component transfer functions, including conversion factors as necessary (and oriented so that, if possible, the units would cancel):

$$\frac{0.025\ \cancel{\text{in.}}}{\text{track}} \times \frac{20\ \cancel{\text{threads}}}{\cancel{\text{in.}}} \times \frac{360\cancel{°}}{\cancel{\text{thread}}} \times \frac{1\ \text{step}}{30\cancel{°}} = \frac{6\ \text{steps}}{\text{track}}$$

Thus, the stepper motor must take six steps to advance one track on the disk.

The program must first read the switches, then calculate the required number of steps, and, finally, output the step command pulses, one by one, to the UCN-5804B. The direction bit must also be read and passed along to the UCN-5804B. Figure 8.34 shows a simplified flowchart of the program.

The next step is to translate the flowchart into a BASIC program. The complete program, along with line-by-line explanations, is given in Table 8.1. With BASIC we can only input or output 8 bits at a time. The input is the 8 bits from the switches. For output, only 2 of the 8 bits are used: the least significant bit (LSB) (D_0) for the step input pulse and the most significant bit (MSB) (D_7) for the direction command. The

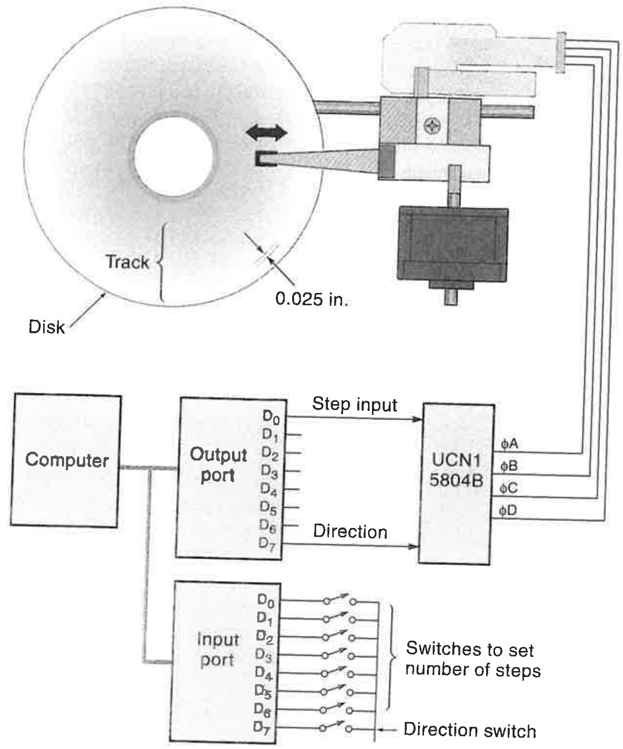


Figure 8.33
The hardware setup for the stepper motor example.

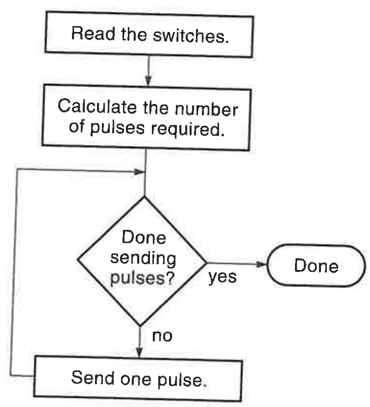


Figure 8.34
Flowchart for a program to drive a stepper motor.

TABLE 8.1 BASIC Program to Control a Disk Drive Head

	Instruction	Explanation
10	SW = INP(209)	Input 8 bits of switch data.
20	IF SW < 128 THEN	If MSB = 0, then the direction bit = 0.
	DIR = 0	
	ELSE	
	DIR = 128	Otherwise, set direction bit = 1 (10000000 = 128)
	SW = SW − 128	and remove the MSB to leave the number of tracks.
	END IF	
30	PUL = SW*6	Calculate the number of pulses needed at 6 pulses per track.
40	REM***** SEND PULSES****	
50	FOR I = 1 TO PUL	Prepare to send PUL (number of pulses).
60	HIGH = 1 + DIR	Make the LSB a 1 and add a direction bit.
70	OUT 208, 1	Send out a "high" for the step pulse.
80	FOR J=1 TO 100	This is the time-delay loop for the pulse width.
90	NEXT J	
100	LOW = 0 + DIR	Make the LSB a 0 and add a direction bit.
110	OUT 208, 0	Send out a "low" to define the end of the pulse.
120	FOR J = 1 TO 100	This is the time-delay loop for the pulse being "low."
130	NEXT J	
140	NEXT I	Go back and send the next pulse.
END		

step input pulse is created by making the output (D_0) go "high" for a period of time and then bringing it "low" for a period of time. The direction bit, brought in as the MSB from the switches, is simply passed on through by the program and sent out as the MSB (actually it is stripped off the input data and added back later to the output word). The parallel I/O port addresses would, of course, depend on the system being used; in this case, the input switch port is 209 (decimal), and the output port to the motor has an address of 208 (decimal). ♦

SUMMARY

A stepper motor, a unique type of DC motor, rotates in fixed steps of a certain number of degrees: 30°, 15°, 1.8°, and so on. The rotor (the part that moves) is made of permanent magnets or iron and contains no coils and therefore has no brushes. Surrounding the rotor is the stator, which contains a series of field pole electromagnets. As the electromagnets are energized one after the other, the rotor is pulled around in a circle. Stepper motors used in position systems are usually operated open-loop, that is, without feedback sensors. The controller will step the motor so many times and expect the motor to be there.

The permanent magnet (PM) stepper motor uses permanent magnets in the rotor. This type of motor has a detent torque—a small magnetic tug that tends to keep it in the last position stepped to, even when the power is removed. PM motors have good torque capability but cannot take very small steps. The field coils of the PM motor can be driven in the two-phase or four-phase mode. The two-phase mode requires polarity reversals. Four-phase operation does not require polarity reversals and the timing is more straightforward.

The variable reluctance (VR) stepper motor uses a toothed iron wheel for the rotor (instead of permanent magnets). This allows VR motors to take smaller steps, but they are weaker and have no detent torque.

The hybrid stepper motor combines the features of both the PM and VR stepper motors and is the type in most common use today. Hybrid motors can take small steps (typically 1.8°) and have a detent torque. The internal construction of the hybrid motor is more complicated than the PM or VR motors, but the electrical operation is just as simple.

Stepper motors are driven from digital circuits that provide the desired number of stepping pulses (in the correct order) to the field coils. These pulses must usually be amplified with driver transistors (operating as switches) before being applied to the motor coils. ICs are available that can provide the proper sequencing and amplification in one chip.

GLOSSARY

ballast resistor A resistor placed in series with the motor coils to improve the torque at higher stepping rates; it works by reducing the motor-time constant.

bilevel drive A technique that uses two voltages to improve torque at higher stepping rates. A higher voltage is applied to the motor at the beginning of the step, and then a lower voltage is switched in.

bipolar motor A motor that requires polarity reversals for some of the steps; a two-phase motor is bipolar.

constant current chopper drive A drive circuit for stepper motors that uses PWM techniques to maintain a constant average current at all speeds.

cumulative error Error that accumulates; for example, a cumulative error of 1° per revolution means that the measurement error would be 5° after five revolutions.

detent torque A magnetic tug that keeps the rotor from turning even when the power is off; also called *residual torque*.

dynamic torque The motor torque available to rotate the load under normal conditions.

eight-step drive A two- or four-phase motor being driven in half-steps; the sequencing pattern has eight steps.

four-phase stepper motor A motor with four separate field circuits; this motor does not require polarity reversals to operate and hence is unipolar.

four-step drive The standard operating mode for two- or four-phase stepper motors taking full steps; the sequencing pattern has four states.

half-steps By alternating the standard mode with dual excitation mode, the angle of step will be half of what it normally is.

holding torque The motor torque available to keep the shaft from rotating when the motor is stopped but with the last field coil still energized.

hybrid stepper motor A motor that combines the features of the PM and VR stepper motors—that is, it can take small steps and has a detent torque.

L/R drive A stepper motor driver circuit that uses ballast resistors in series with the motor coils to increase torque at higher stepping rates.

microstepping A technique that allows a regular stepper motor to take fractional steps; it works by energizing two adjacent poles at different voltages and by balancing the rotor between.

permanent magnet (PM) stepper motor A motor that uses one or more permanent magnets for the rotor; this motor has a detent torque.

phase The number of separate field winding circuits.

rotor The internal part of the stepper motor that rotates.

single-step mode Operating the motor at a slow enough rate so that it can be stopped after any step without overshooting.

slew mode Stepping the motor at a faster rate than the single-step mode; used to move to a new position quickly. The motor will overshoot if the speed is not ramped up or down slowly.

stalling A situation wherein the motor cannot rotate because the load torque is too great.

stator The part of a stepper motor that surrounds the rotor and consists of field poles (electromagnets).

stepper motor A motor that rotates in steps of a fixed number of degrees each time it is activated.

three-phase stepper motor A motor with three separate sets of field coils; usually found with VR motors.

two-phase stepper motor A motor with two field circuits. This motor requires polarity reversals to operate and hence is bipolar.

unipolar motor A motor that does not require polarity reversals. A four-phase motor is unipolar.

variable reluctance (VR) stepper motor A motor that uses a toothed iron wheel for the rotor and consequently can take smaller steps but has no detent torque.

EXERCISES

Section 8.1

1. A 15° stepper motor is commanded to go 100 steps CW and 30 steps CCW from a reference point. What is its final angle?
2. A 7.5° stepper motor is commanded to go 50 steps CCW, 27 steps CW, and 35 steps CCW again. What is its final angle (referenced from its original position)?
3. Why can a stepper motor be operated open-loop in a control system?
4. What is the detent (or residual) torque in a PM stepper motor, and what causes it?
5. A stepper motor is being used as a crane motor in an expensive toy. The motor has the following properties: Holding torque = 35 in. · oz, dynamic torque = 20 in. · oz, and detent torque = 5 in. · oz. The motor turns a 1.5-in. diameter pulley around which a string is wound. How much weight can the crane lift? How much weight can the crane continue to support with the power on; with the power off?
6. List the stepping sequence you would use to make the two-phase motor of Figure 8.8 rotate CW.
7. List the stepping sequence you would use to make the two-phase motor of Figure 8.9 operate as an eight-step drive (CCW).
8. List the stepping sequence you would use to make the four-phase motor of Figure 8.11(a) operate as an eight-step drive (CCW).

Section 8.2

9. Explain the principle of operation of a VR stepper motor.
10. Does a VR stepper motor have a detent torque? Explain.

Section 8.3

11. Explain the principle of operation of a hybrid stepper motor.
12. What are the advantages of the hybrid stepper motor?

Section 8.4

13. A 5-V stepper motor is to be microstepped with one-tenth steps. List the voltage table required for this [similar to Figure 8.27(b)].
14. What is the purpose of a ballast resistor on a stepper motor drive, and how does it work?
15. What is the purpose of bilevel drive, and how does it work?
16. How does a chopper drive improve torque at higher stepping rates?

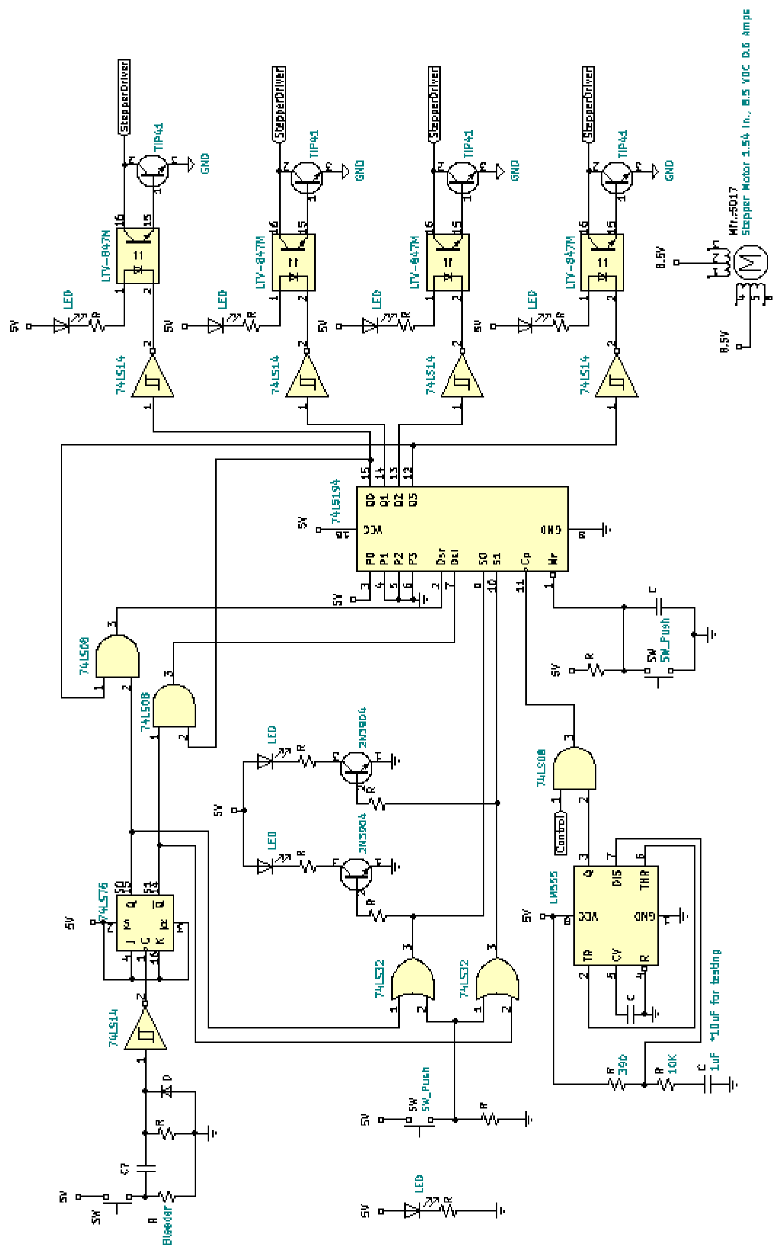
Section 8.5

17. A 1.8° stepper motor turns a leadscrew that has 24 threads per inch.
    a. How many steps will it take to advance the leadscrew 1.25 in.?
    b. What is the linear distance the leadscrew advances for each step?
18. A 7.5° stepper motor (four phase), controlled by a computer, is used to position a telescope through a gear train. The telescope must be positioned to within 0.01°. A front panel has toggle switches that are used to specify how far the telescope is to move (LSB = 0.01°). The total range of the telescope is 0–60°.
    a. How many toggle switches would be required?
    b. What gear ratio would you specify?
    c. Draw a block diagram of the system, showing all parts of the system and specifying the gear ratio.

14.2 Stepper Motor Circuit Analysis

14.2.1 Stepper Motor Schematic

RCET0251 MOTORS, STEPPER TIM LEISHMAN



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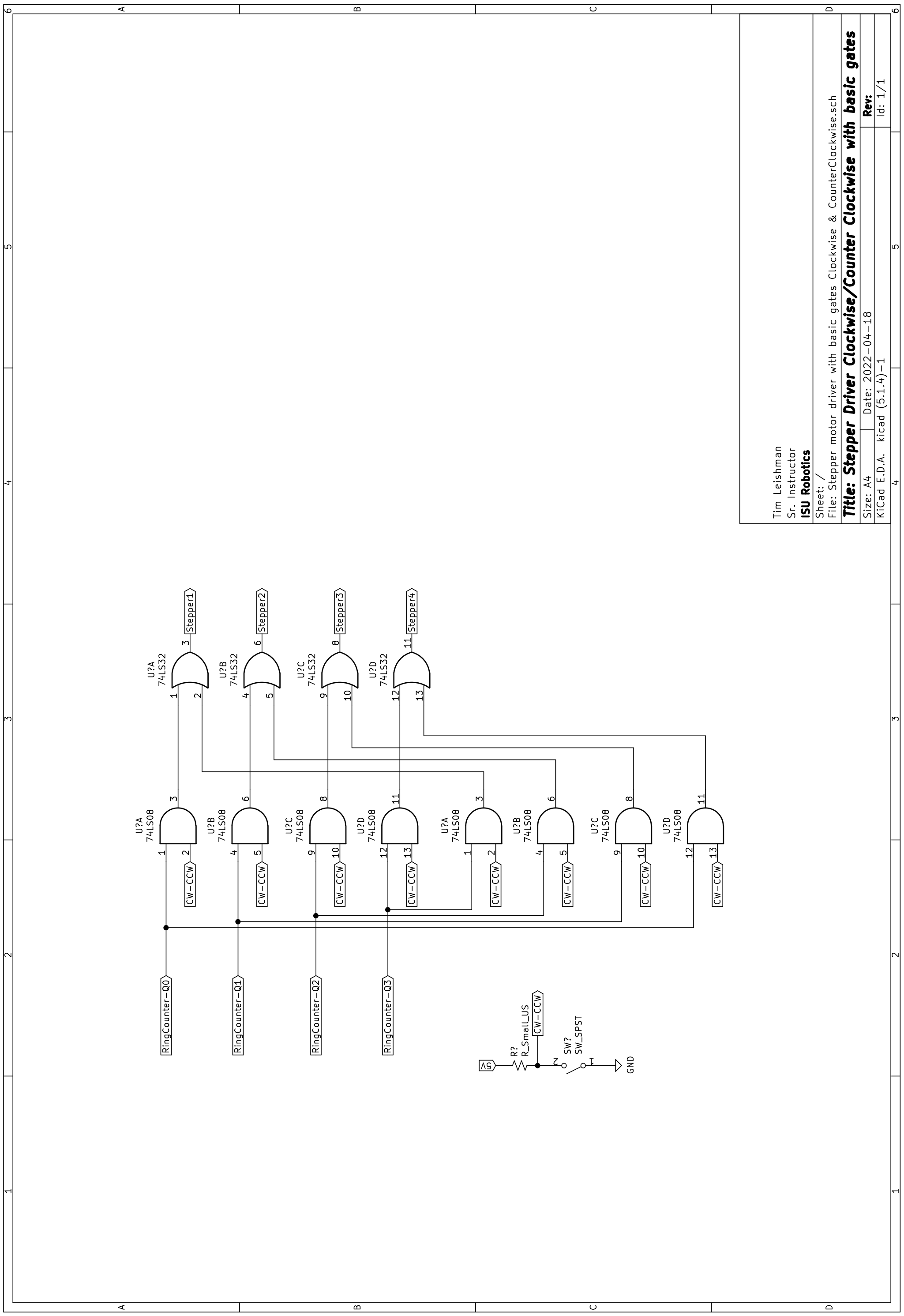
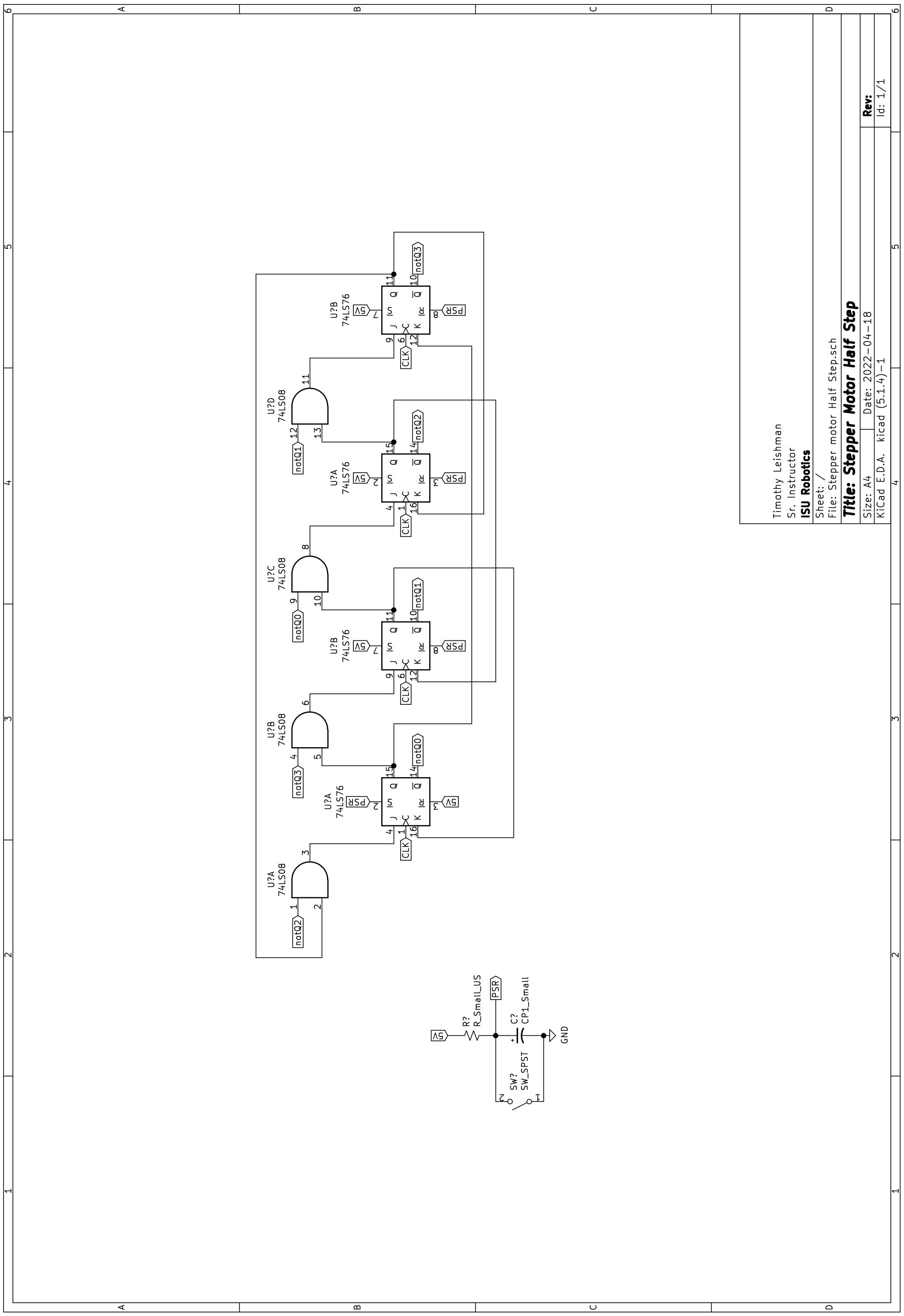
14.2.2 Direction Control Schematic



14.2.3 Direction Control Schematic



Week 15

Review

15.1 RCD Circuit Analysis

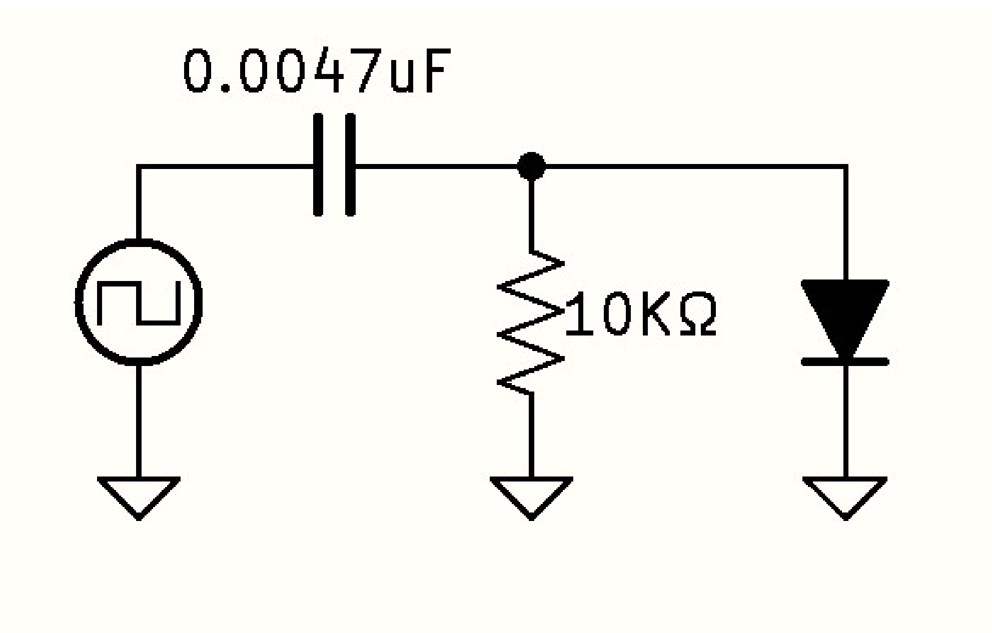


Figure 15.1: RCD Circuit

15.1.1 Circuit Parameters

- The generator square wave voltage is 0 to 8V.
- The generator frequency is 1Khz.

15.1.2 1Khz Square-wave Circuit Analysis

1. Plot the Generator Voltage.
   (a) $Vgen_{PW} = \frac{1}{1Khz} = 1mS$
   (b) $Vgen_{Ton/Toff} = \frac{Vgen_{PW}}{2} = 500uS$
2. Analyze the circuit as though the diode has been removed, and find the resistor voltage.
   (a) $\tau = RC = (10K\Omega Rgen)(0.0047\mu F) = (10.05K\Omega)(0.0047\mu F) = 47.235\mu Sec$
   (b) Review RC Circuits Capacitor % of Charge vs. Tau, 2.1 on page 51.
      - $1\tau = 63.212\%$, data points $(1 \times 47.235\mu Sec, 0.632 \times 8)$
      - $2\tau = 86.466\%$, data points $(2 \times 47.235\mu Sec, 0.865 \times 8)$
      - $3\tau = 95.021\%$, data points $(3 \times 47.235\mu Sec, 0.95 \times 8)$
      - $4\tau = 98.168\%$, data points $(4 \times 47.235\mu Sec, 0.982 \times 8)$
      - $5\tau = 99.326\%$, data points $(5 \times 47.235\mu Sec, 0.993 \times 8)$

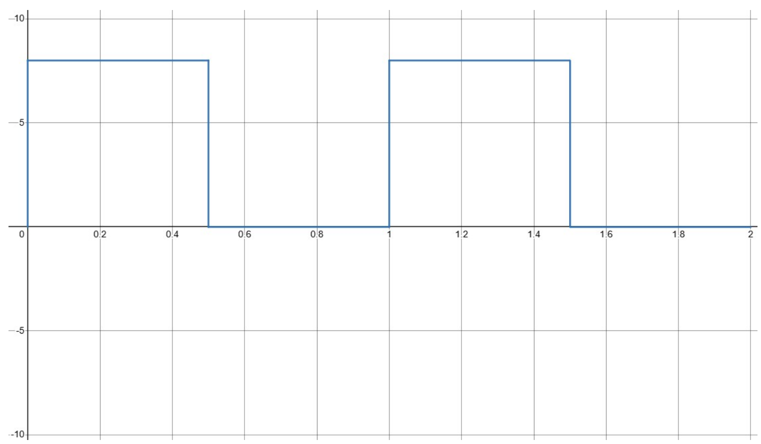


Figure 15.2: Calculated Generator Voltage Waveform without diode

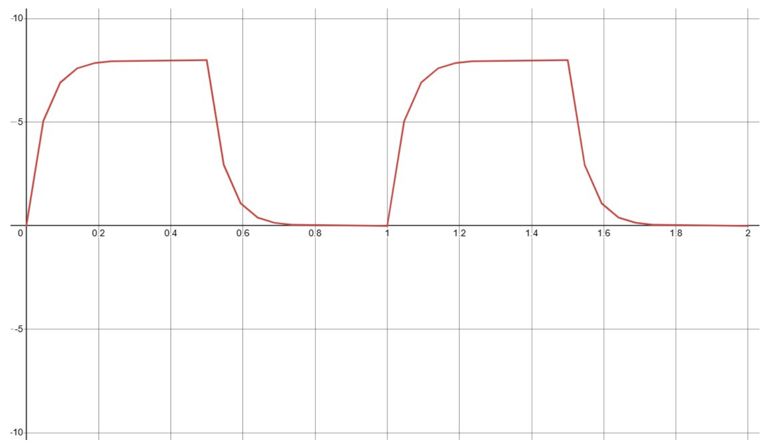


Figure 15.3: Calculated Capacitor Voltage Waveform without diode

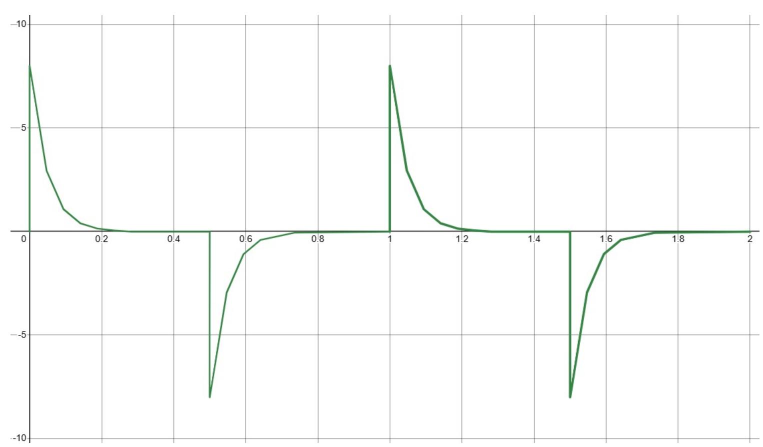


Figure 15.4: Calculated Resistor Voltage Waveform without diode

3. The resistor is in parallel with the diode, and evaluating the resistor voltage, we can see when the diode will become forward biased.

   (a) At time 0 Vr attempts to go to 8V. This will forward-bias the diode.

   (b) Once the diode is forward-biased, the $10K\Omega$ resistor is bypassed and the current path is now through the diode. This means that Rgen and the diode resistance now provide the circuit resistance for the capacitor charge.

   - $R_{Gen} = 50\Omega$
   - $R_{Diode} = \frac{V_{diode}}{I_{diode}}$
     - The maximum diode current will be when the diode is initially forward biased. As the capacitor charges, the diode current will decrease, and once the capacitor is fully charged, the diode current will be zero.
     - The average diode resistance will be equal to the voltage of the diode divided by the average diode current.
     - Initial max current $I_{max} = \frac{V_{gen}-V_{diode}}{R_{Gen}} = \frac{8-0.7}{50\Omega} = \frac{7.3V}{50\Omega} = 146mA$
     - Average diode current $I_{diode} = \frac{I_{max}}{2} = \frac{146mA}{2} = 73mA$
     - $R_{Diode} = \frac{V_{diode}}{I_{diode}} = \frac{0.7V}{73mA} = 9.589\Omega$

   (c) How long will it take for the capacitor to charge to $V_{Gen} - V_{diode}$?

   - While the diode is forward biased:
     - $\tau = C \times (R_{gen} + R_{diode}) = 0.0047\mu F \times (50\Omega + 9.589\Omega) = 0.0047\mu F \times 59.9589\Omega = 281.807nSec$
     - $5\tau = 1.409\mu Sec$

4. It takes $1.409\mu Sec$ for the capacitor to charge to 7.3 volts. Once the capacitor hits 7.3 volts, the diode is no longer forward-biased, the $10K\Omega$ resistor completes the current path, and the circuit RC tau returns to the previously calculated $\tau = 47.235\mu Sec$ for the remaining charge time.

   (a) Remaining Charge Time, Cap will charge from 7.3V to 8V.

   - at time equals $1.409\mu Sec$, $V_C = 7.3v$
   - at time $t = 1.409\mu Sec + 47.235\mu Sec$, V_C will charge an additional 63.212% of the remaining 0.7 volts.
   -
   - at time $t_1 = 1.409\mu Sec + 47.235\mu Sec = 48.644\mu Sec$ $V_C = 7.3V + (0.63212 \times 0.7) = 7.3V + 0.44248 = 7.742V$
   - at time $t_2 = 1.409\mu Sec + 2 \times 47.235\mu Sec = 95.879\mu Sec$ $V_C = 7.3V + (0.86466 \times 0.7) = 7.3V + 0.605 = 7.905V$
   - at time $t_3 = 1.409\mu Sec + 3 \times 47.235\mu Sec = 143.114\mu Sec$ $V_C = 7.3V + (0.95021 \times 0.7) = 7.3V + 0.665 = 7.965V$
   - at time $t_4 = 1.409\mu Sec + 4 \times 47.235\mu Sec = 190.349\mu Sec$ $V_C = 7.3V + (0.98168 \times 0.7) = 7.3V + 0.687 = 7.987V$

- at time $t_5 = 1.409\mu Sec + 5 \times 47.235\mu Sec = 237.584\mu Sec$ $V_C = 7.3V + (0.99326 \times 0.7) = 7.3V + 0.687 = 7.995V$

(b) For the remaining cycle, the diode will remain off and the waveform will remain as previously calculated.

5. The resistor waveform voltage, once the diode is off, can be derived from the capacitor voltage.

   (a) $V_R = V_{Gen} - V_C$

6. Generator Loading, as noted previously, when the diode is initially forward biased, the circuit resistance is equal to $R_{gen} + R_{diode}$ $50\Omega + 9.589\Omega$. Meaning that we would expect to observe generator loading when the diode is forward biased and for approximately $1.409\mu Sec$.

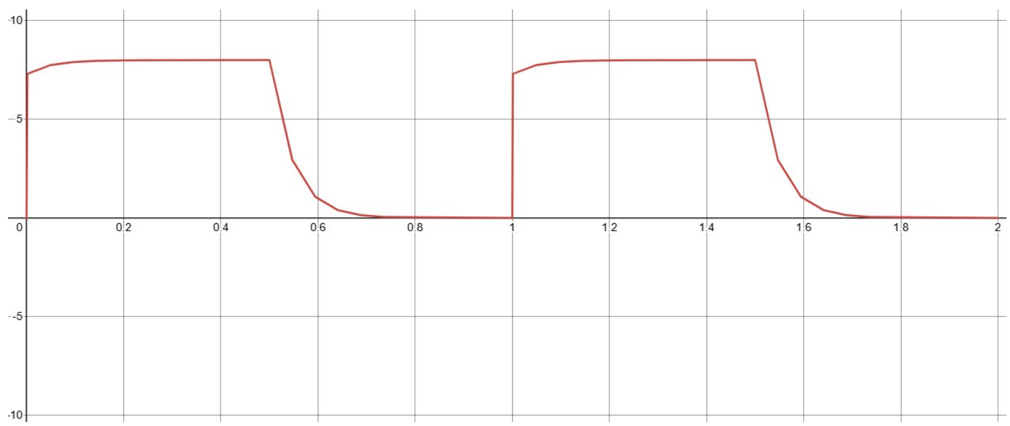


Figure 15.5: Calculated Capacitor Voltage Waveform with diode

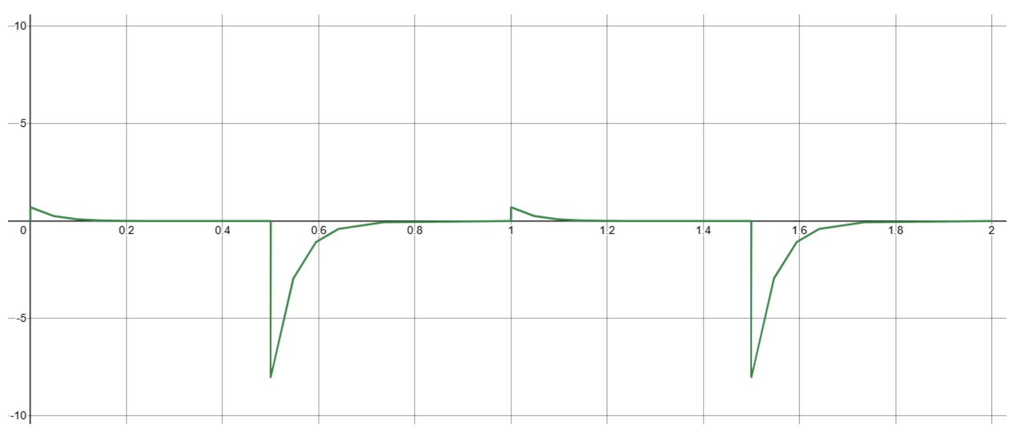


Figure 15.6: Calculated Resistor Voltage Waveform with diode

15.1.3 1Khz Square Wave Measured Waveforms

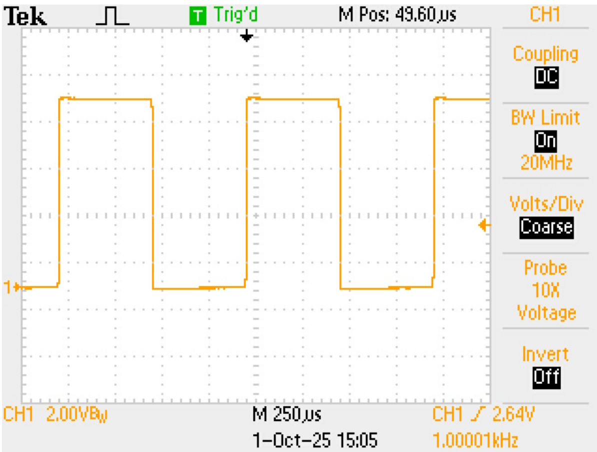


Figure 15.7: 1Khz Measured Vgen without diode

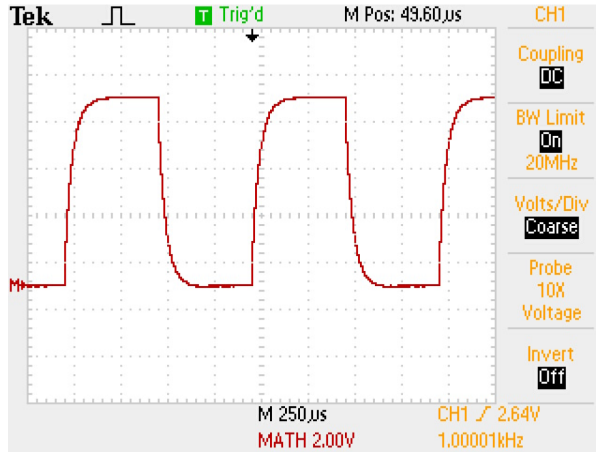


Figure 15.8: 1Khz Measured VC without diode

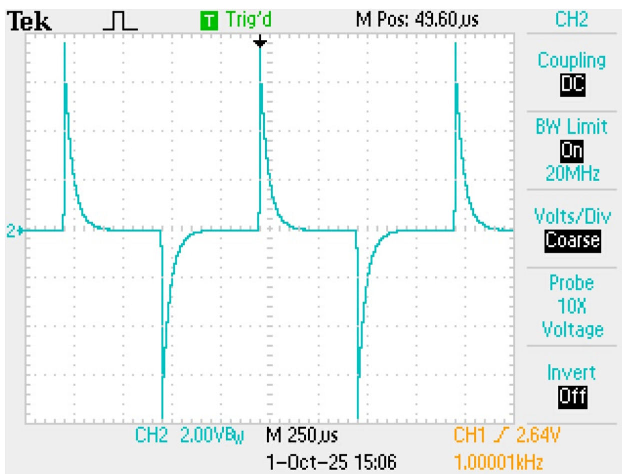


Figure 15.9: 1Khz Measured VR Waveform without diode

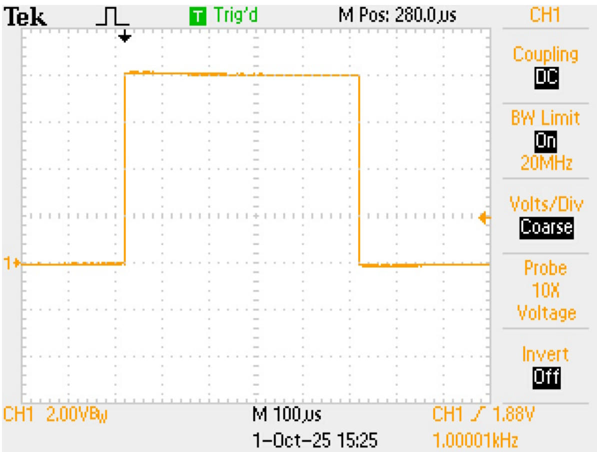


Figure 15.10: 1Khz Measured VGen with diode

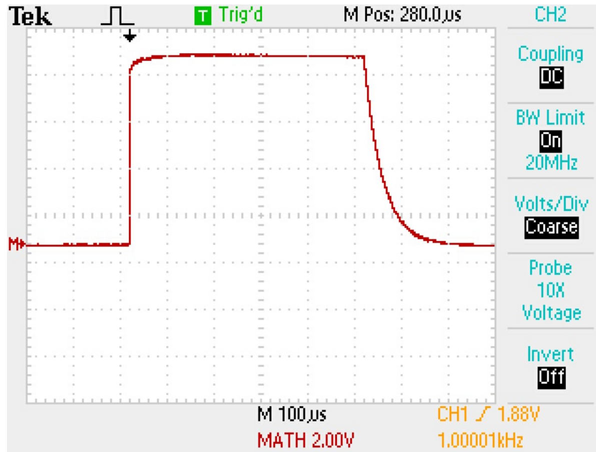


Figure 15.11: 1Khz Measured VC with diode

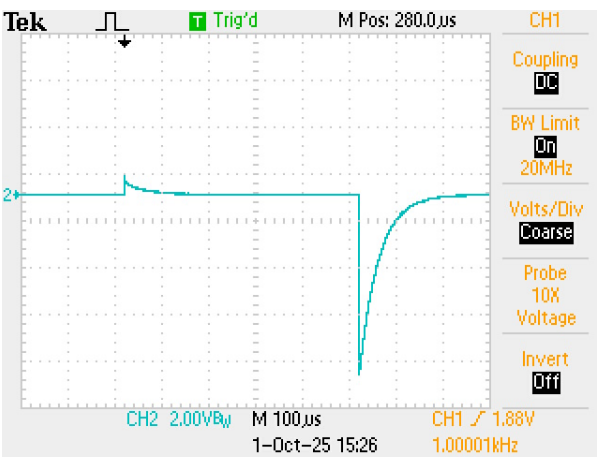


Figure 15.12: 1Khz Measured VR with diode

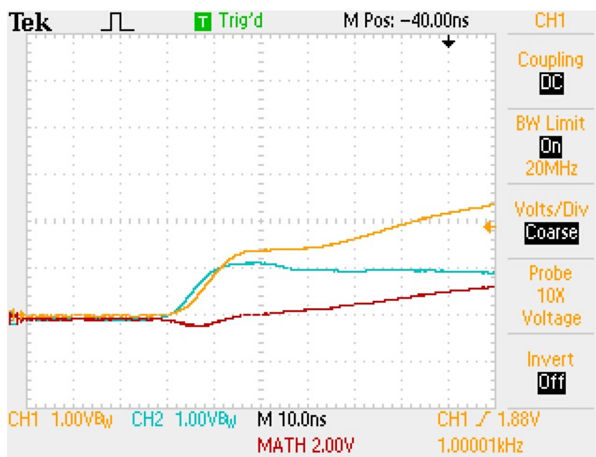


Figure 15.13: 1Khz 10nSec per div with diode

15.1.4 10Khz Square Wave Circuit Analysis

- See Circuit Figure 15.1
- The generator square-wave voltage is 0 to 8V.
- The generator frequency is 10Khz.

1. Plot the Generator Voltage.
   (a) $Vgen_{PW} = \frac{1}{10Khz} = 100\mu Sec$
   (b) $Vgen_{Ton/Toff} = \frac{Vgen_{PW}}{2} = 50uSec$
2. Analyze the circuit as though the diode has been removed, and find the resistor voltage.
   (a) $\tau = RC = (10K\Omega Rgen)(0.0047\mu F) = (10.05K\Omega)(0.0047\mu F) = 47.235\mu Sec$
   (b) Calculate the Capacitor Voltage. Notice that the $PW \approx RC(\tau)$. This means the capacitor will not have enough time to fully charge or discharge. Review cycles to stabilization Section 2.9.1 on page 53. However, we can use logic to predict the capacitor charge waveform. Because there is not enough time for the cap to charge and that the generator is at 50% duty cycle, the capacitor voltage waveform will center at the average voltage of 4V. Knowing that we have $\approx 1\ \tau$ to charge and discharge, we can predict that the stablized max and min voltage will be plus and minus 63% of 4V.
      - $V_{Max} = \frac{V_{gen+}}{1+e^{\frac{-PW}{RC}}}$
        - $V_{Max} = \frac{8V}{1+e^{\frac{-50\mu sec}{47.235\mu Sec}}} = 5.939V$
      - $V_{Min} = V_{gen+} - V_{Max}$
        - $V_{Min} = 8V - 5.939V = 2.061V$
   (c) Calculate the Resistor Voltage using Kirchhoff.

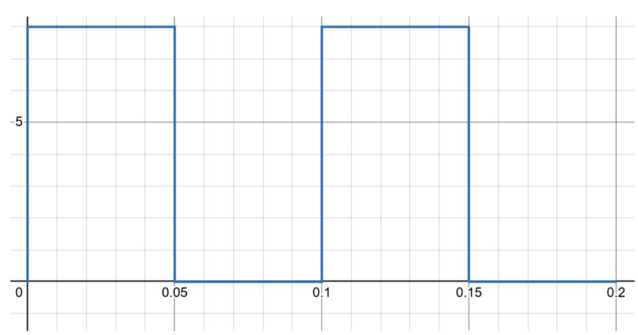


Figure 15.14: 10Khz Calculated Generator Voltage Waveform without diode

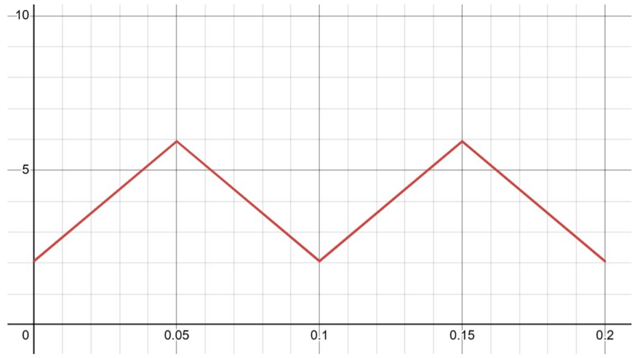


Figure 15.15: 10Khz Calculated Capacitor Voltage Waveform without diode

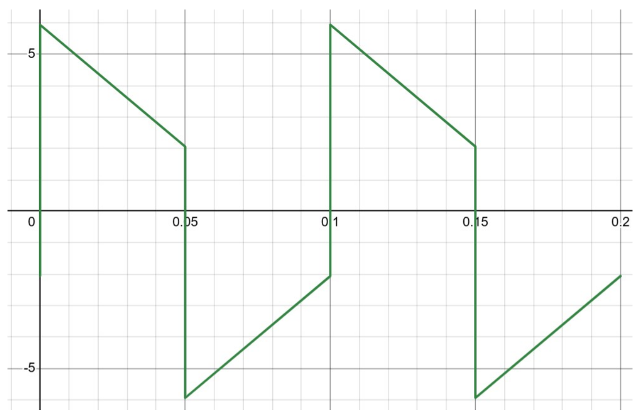


Figure 15.16: 10Khz Calculated Resistor Voltage Waveform without diode

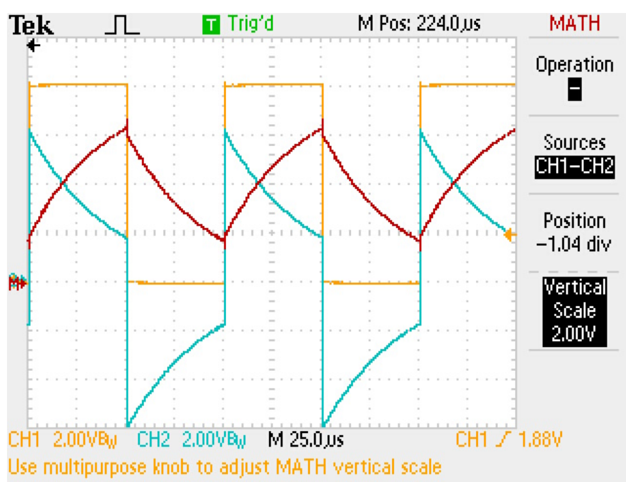


Figure 15.17: 10Khz measured VGen, VR, and VC Waveforms without diode

3. Analyze Circuit with Diode.

   - As the generator goes positive, we can see that the diode will become forward biased.
   - The diode will remain forward-biased until the capacitor charges to 7.3V.
   - When the diode is forward biased, 5tau of the circuit becomes $1.409\mu Sec$ (previously calculated in the 1Khz section).
   - After $1.409\mu Sec$, the diode is off and the capacitor will continue to charge for the remainder of the pulse width through the $10K\Omega$ resistor. The tau is now $47.235\mu Sec$ (previously calculated in the 1Khz section). However, the remaining time of the PW is $50\mu Sec - 1.409\mu Sec = 48.591\mu sec$. The capacitor has $\approx 1\tau$ to charge or 63% of 0.7V and the capacitor voltage will be $Vc = 7.3V + (.63 \times 0.7V) = 7.3V + 0.441 = 7.741V$
   - As the generator voltage goes to zero (PW), the capacitor will have $50\mu Sec$ to discharge.

     $Vc = Vfin - (Vfin - Vin)e^{\frac{-t}{\tau}}$

     $Vc = 0 - (0 - 7.741V)e^{\frac{-50\mu Sec}{47.235\mu Sec}}$

     $Vc = (7.741V)e^{\frac{-50\mu Sec}{47.235\mu Sec}}$

     $Vc = 2.686V$
   - The resistor voltage waveform can be calculated using Kirchhoff.

Calculated Waveforms

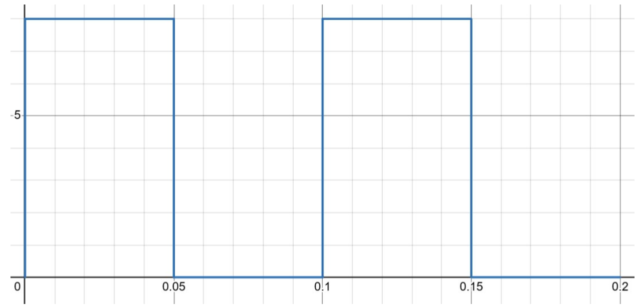


Figure 15.18: 10Khz Calculated Generator Voltage Waveform with diode

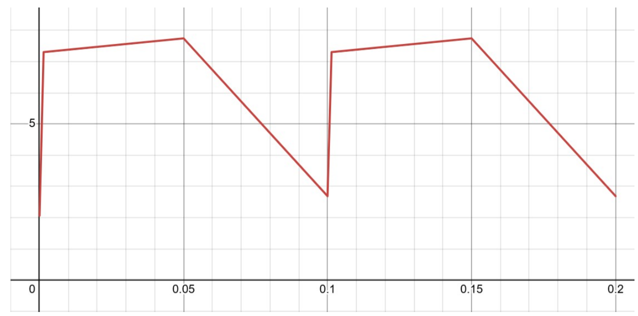


Figure 15.19: 10Khz Calculated Capacitor Voltage Waveform with diode

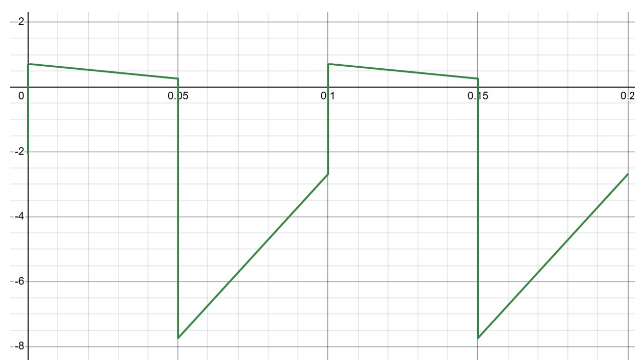


Figure 15.20: 10Khz Calculated Resistor Voltage Waveform with diode

Measured Waveforms:

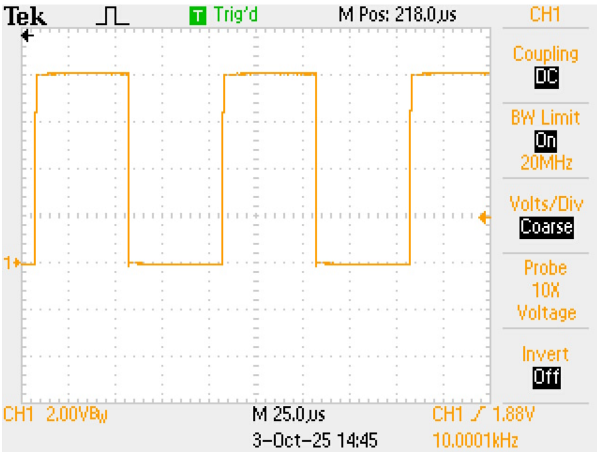


Figure 15.21: 10Khz Measured Vgen with diode

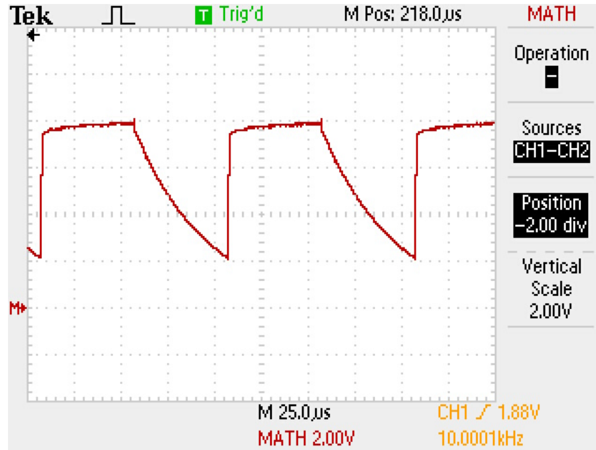


Figure 15.22: 10Khz Measured VC with diode

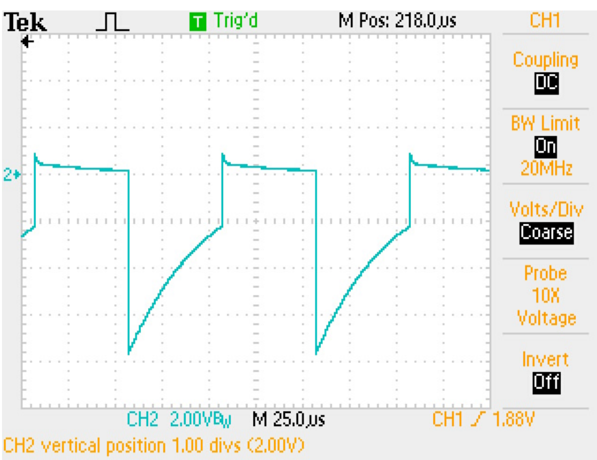


Figure 15.23: 10Khz Measured VR with diode

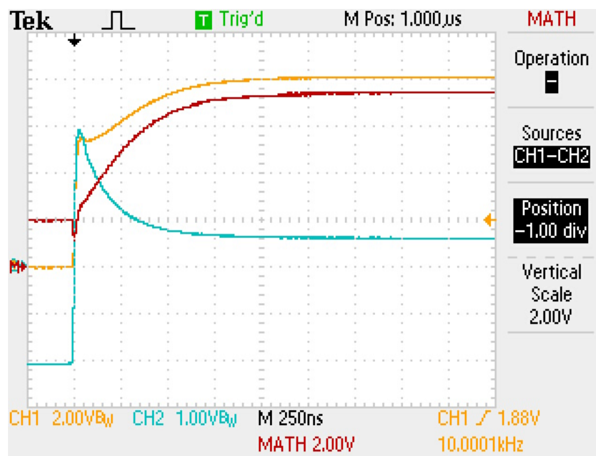


Figure 15.24: 10Khz Measured 250nSec per div with diode

15.1.5 Sine Wave Analysis

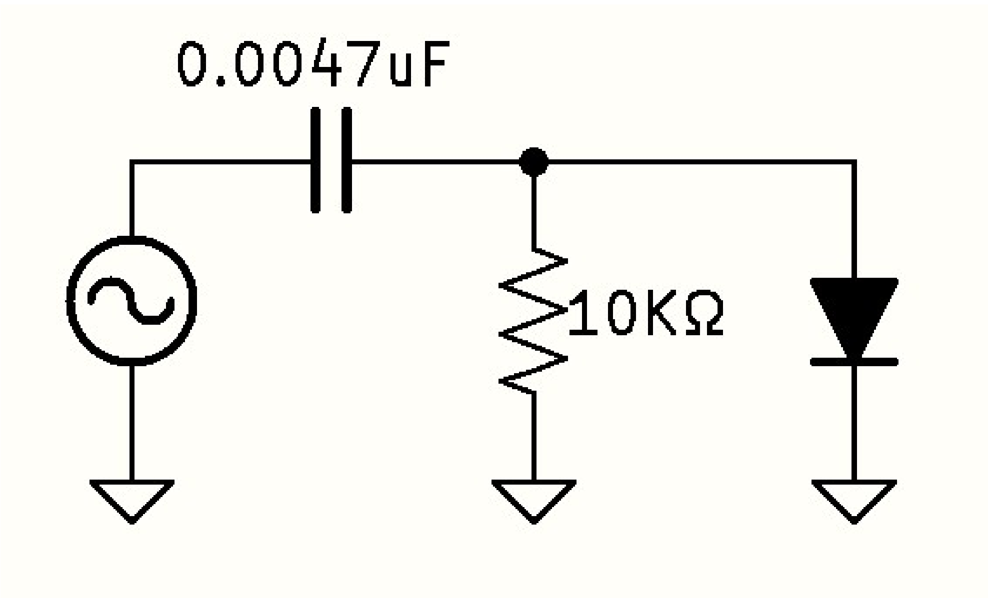


Figure 15.25: RCD Circuit

1Khz Circuit Analysis without Diode

1. Calculate X_C
   - $X_C = \frac{1}{2\pi FC} = \frac{1}{2\pi 1Khz 0.0047\mu F} = 33.862K\Omega,\ -90°$
2. Calculate Z_T
   - $Z_T = \sqrt{(R)^2 + (X_C)^2} = \sqrt{(10K)^2 + (33.862K)^2} = 35.307K\Omega$
   - $\theta = tan^{-1}(\frac{O}{A}) = tan^{-1}(\frac{-33.862K}{10K}) = -73.547°$
3. Calculate I_T
   - $I_T = \frac{V_T}{Z_T} = \frac{8vp}{35.307K\Omega, -73.547°} = 226.583\mu Ap, 73.547°$
4. Calculate V_R
   - $V_R = I_T \times R = 226.583\mu Ap \times 10K\Omega = 2.266vp, 73.547°$
5. Calculate V_C
   - $V_C = I_T \times X_C = 226.583\mu Ap \times 33.862K\Omega = 7.673vp, -16.453°$

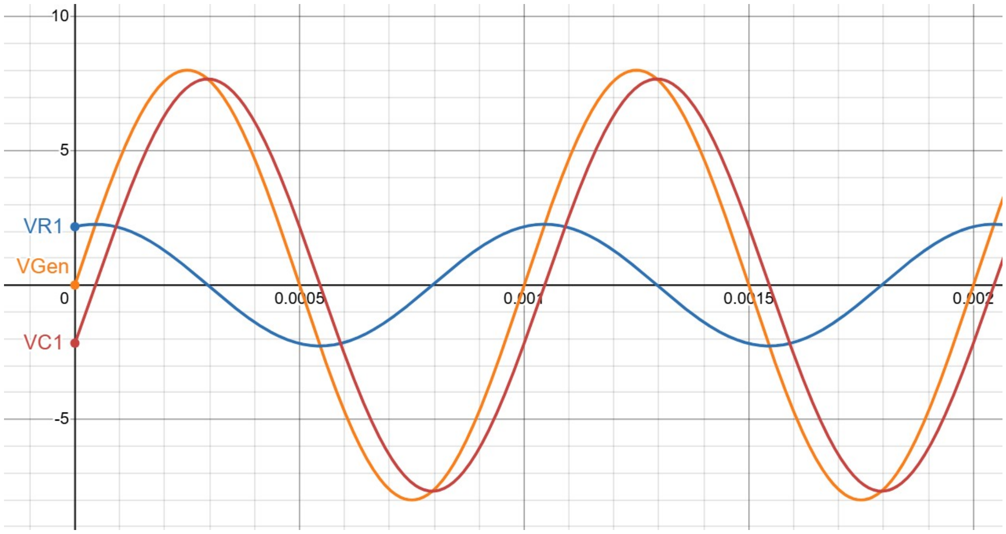


Figure 15.26: RCD 1Khz Sinewave Predicted Waveforms without Diode

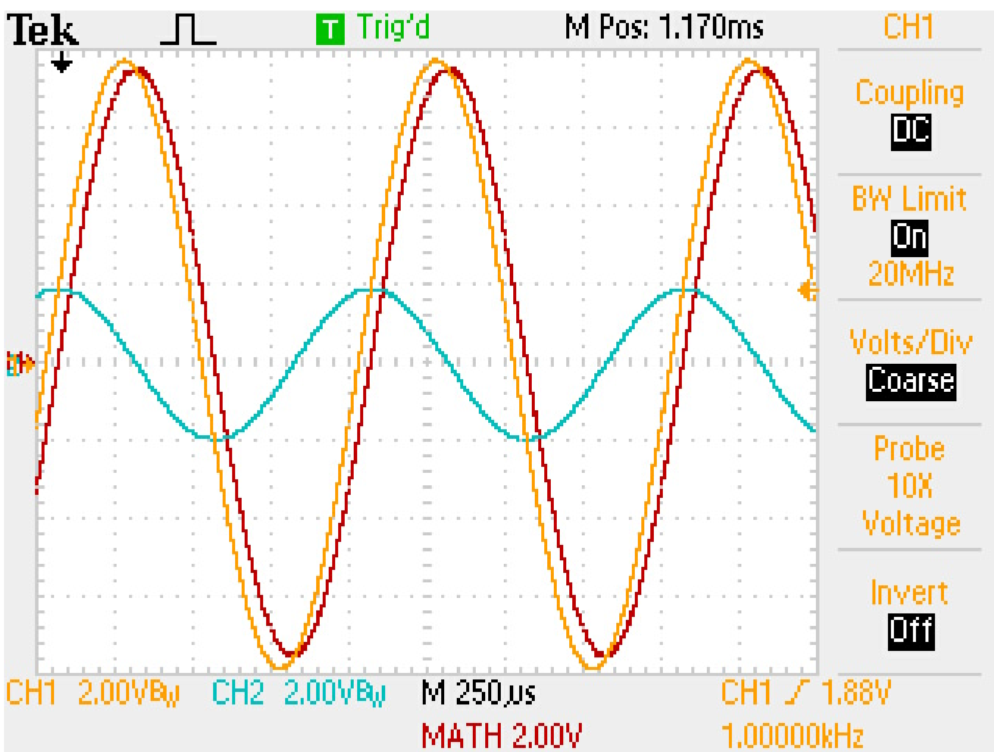


Figure 15.27: Measured RCD 1Khz Sinewave Waveforms without Diode

1Khz Circuit Analysis with Diode

The diode will forward-bias as the resistor voltage approaches the positive 2.266vp. Once the diode is forward-biased, the circuit resistance is 59.9589Ω as previously calculated in step 3c on page 428. We can now find Z_T while the diode is forward-biased, $X_C = 33.862K\Omega$ as previously calculated 1 on page 437.

1. Calculate Z_T
   - $Z_T = \sqrt{(R)^2 + (X_C)^2} = \sqrt{(59.959)^2 + (33.862K)^2} = 33.862K\Omega$
   - $\theta = tan^{-1}(\frac{O}{A}) = tan^{-1}(\frac{-33.862K}{59.959}) = -89.9°$
2. Calculate I_T
   - $I_T = \frac{V_T}{Z_T} = \frac{8vp}{33.862K\Omega, -89.9°} = 236.253\mu Ap, 89.9°$
3. Calculate V_R
   - $V_R = I_T \times R = 236.253\mu Ap \times 59.959\Omega = 14.165mVp, 89.9°$
4. Calculate V_C
   - $V_C = I_T \times X_C = 236.253\mu Ap \times 33.862K\Omega = 8vp, -0.1°$

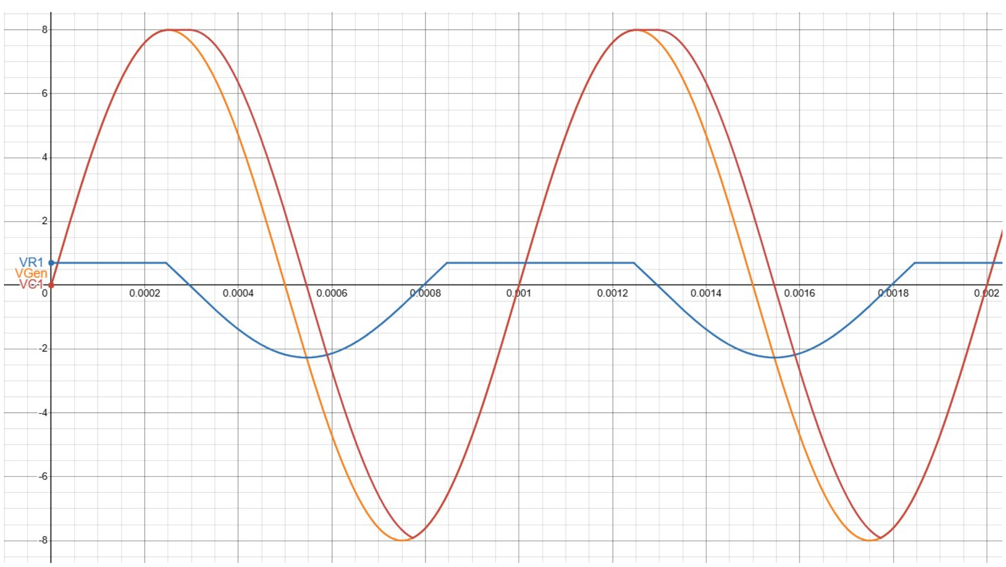


Figure 15.28: RCD 1Khz Sine wave Predicted Waveforms with Diode

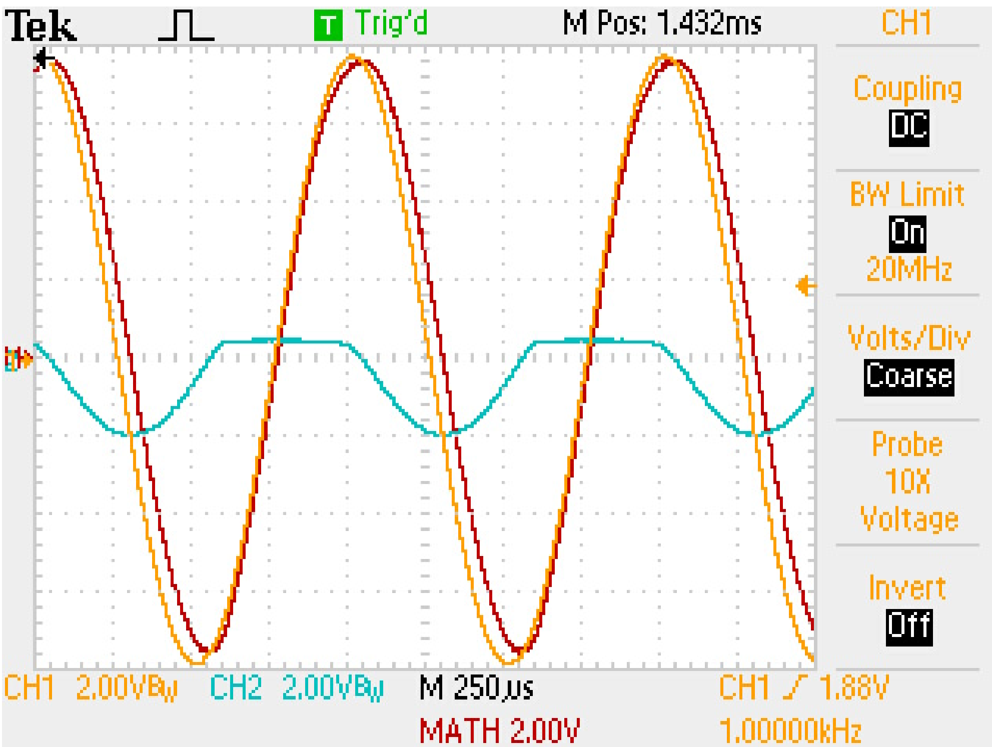


Figure 15.29: Measured RCD 1Khz Sine wave Waveforms with Diode

10Khz Analysis with diode removed

1. Calculate X_C
   - $X_C = \frac{1}{2\pi FC} = \frac{1}{2\pi 10Khz 0.0047\mu F} = 3.386K\Omega,\ -90°$
2. Calculate Z_T
   - $Z_T = \sqrt{(R)^2 + (X_C)^2} = \sqrt{(10K)^2 + (3.386K)^2} = 10.558K\Omega$
   - $\theta = tan^{-1}(\frac{O}{A}) = tan^{-1}(\frac{-3.386K}{10K}) = -18.706°$
3. Calculate I_T
   - $I_T = \frac{V_T}{Z_T} = \frac{8vp}{10.558K\Omega, -18.706°} = 757.719\mu Ap, 18.706°$
4. Calculate V_R
   - $V_R = I_T \times R = 757.719\mu Ap \times 10K\Omega = 7.577vp, 18.706°$
5. Calculate V_C
   - $V_C = I_T \times X_C = 757.719\mu Ap \times 3.3862K\Omega = 2.565vp, -71.294°$

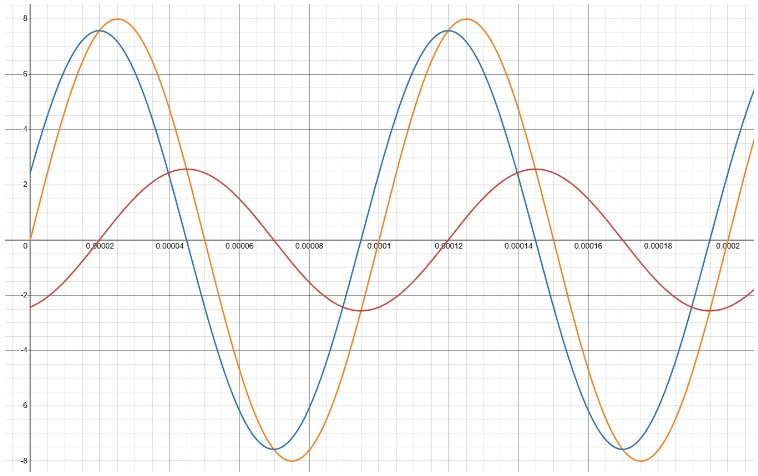


Figure 15.30: RCD 10Khz Sinewave Predicted Waveforms without Diode

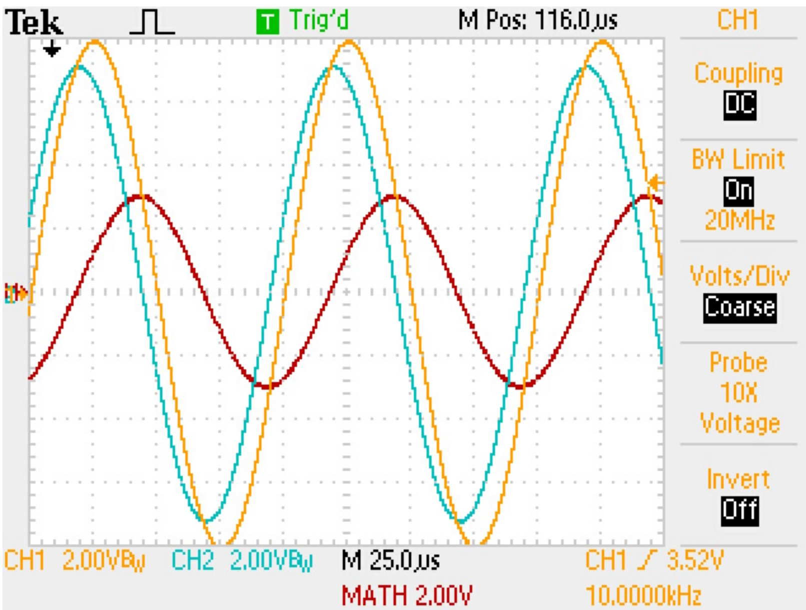


Figure 15.31: RCD 10Khz Sinewave Measured Waveforms without Diode

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