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Diseño de la instrumentación y
protección de restaurador
dinámico de voltaje.

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21/04/2015

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CAPITULO I

1.1- Introducción

La electrónica en la actualidad es un pilar fundamental para la humanidad, ya que está implícita en casi todas las actividades que desarrollamos, incluso en las sociales, sin embargo la electrónica desde sus orígenes hasta hoy en día ha jugado un papel muy importante y casi fundamental en el desarrollo de las empresas, y por lo tanto en la economía. Pero, como se ha logrado esto? ; simplemente por la evolución de la misma electrónica desde sus orígenes, hasta llegar a los sistemas de control y automatización más complejos que en un nivel industrial, nos sirven en diversas áreas y tienen muchas ventajas como mejorar un proceso, o evitar accidentes, mejor producción , optimización peor sobre todo la seguridad que nos proporciona la automatización.

Sin embargo con la automatización y la demanda de nuevos sistemas para el control de los equipos electromecánicos con los que se manejaban casi todas las industrias a nivel mundial fue un reto para la electrónica de potencia, por lo que se tuvo que incursionar en nuevos procesos, en la miniaturización de los dispositivos electrónicos y en la optimización de la energía ; sin embargo la energía eléctrica por su propia naturaleza tiende a tener disturbios o armónicos que incluyeron un nuevo reto a los sistemas de automatización y protección, ya que se busca alta calidad en cualquiera de los procesos industriales o incluso domésticos.

Los disturbios eléctricos merman la calidad de la energía eléctrica y se clasifican por su tipo y duración y existen estándares para clasificarlos.

Debido a esto se han buscado técnicas para censar estos disturbios eléctricos para reducir daños al equipo y o al personal que lo opere , así como obtener una mejor calidad y eficiencia de la energía eléctrica, ya que estos armónicos, o disturbios afectan a los sistemas electrónicos sensibles, o controladores; censando en todo esto se desarrollo lo que conocemos coloquialmente como DVR o regulador dinámico de voltaje que es una solución para la mitigación de los disturbios de voltaje tipo SAG y SWELL, y se combina con diversas técnicas y algoritmos varios para la mitigación de estos disturbios eléctricos.

Es por eso que en este proyecto nos enfocaremos a lo que la electrónica ha logrado desarrollar para este tipo de sistemas y se busca implementar un sistema de protección, y como su nombre lo dice nos sirve para proteger o aislar equipos o dispositivos como medida de protección, o en este caso también para poder medir y registrar las distorsiones que ingresan al sistema el cual el DVR se encargara de mitigar, sin embargo para obtener una grafica cuantificable con datos y parámetros que se puedan utilizar para localizar estas distorsiones y corregirlas, ya que a veces es tan rápido el disturbio que el DVR no alcanza a localizarlo y brinca su tiempo de trabajo, por lo que el proceso de localización y discriminación se complica.

El DVR es un dispositivo el cual se ve disminuido o potencializado por el esquema de censado que es al final de cuentas es el que informa al sistema de control de las condiciones en que se está llevando el proceso de corrección o mitigación de disturbios de energía eléctrica, y por supuesto el sistema de protección que protege al DVR de cualquier incidente que pudiera ocurrir en el desarrollo de la operación. Uno de los problemas importantes en la operación del DVR es la capacidad que tiene el DVR para identificar el instante en que el evento ocurre y entra en operación, pero también cuando el evento ya desapareció de la red y el DVR debe dejar de operar, este proceso de discriminación es un proceso difícil de identificar por la velocidad con la que los eventos ocurren o se mezclan con ruido u otras fallas en línea.

1.2- Justificación

Para que el sistema de control de restauración funcione eficientemente, es importante que los dispositivos a utilizar, tales como los sensores sean los más adecuados en relación al tipo, función, dimensionamiento y acondicionamiento de estos, se propone diseñar , seleccionar e implementar los sistemas de censado y protecciones en el DVR de cualquier incidente que pudiera existir en el desarrollo de la operación, todo esto para que se llegue a la corrección de disturbios tipo SAG y SWELL en la red de energía eléctrica, pudiendo discriminar estos eventos en la red y proteger de cualquier falla producida en la operación que se lleve a cabo tanto interno como externo.

1.3-Objetivos

1.3.1 *Objetivo General*

Diseñar e implementar la instrumentación y protección de un DVR, basado en sensores analógicos y digitales emergentes.

1.3.2 *Objetivos Específicos*

1. Definir la instrumentación y el tipo de protección a utilizar.
2. Seleccionar los tipos de sensores electrónicos a utilizar.
3. Dimensionar y acondicionar las características de las protecciones requeridas para una operación segura.
4. Implementar la instrumentación electrónica y protecciones en software para simulación por medio de un microcontrolador.

CAPITULO II

2.1- Alcances y limitaciones

En la realización de este proyecto se pretende implementar un prototipo utilizando sensores específicos para obtener su comportamiento en una grafica o mapeo utilizando un software diseñado para esa tarea, para poder identificar y analizar el comportamiento de la corriente como medio de instrumentación para la protección de un DVR.

Las primeras pruebas que se llevaron a cabo en la realización del proyecto, se llevaron a cabo con dispositivos electrónicos discretos para la simulación de los sensores, posteriormente, se anexaron los sensores con los cuales se va a trabajar en el prototipo final.

Así mismo por la aplicación del prototipo como, una base general la etapa de protección adjunta posterior a la instrumentación será descrita de manera teórica, por los recursos y la extensión en la investigación y el desarrollo del software.

CAPITULO III

3.1- Marco Teórico.

3.1.1 Sensores.

Cualquiera que sea el sistema de medición de energía, debe tener elementos sensores de corriente y de voltaje. El sensor de corriente convierte la magnitud de corriente de un alinea de distribución, a un nivel de corriente o voltaje directamente equivalente. El sensor de corriente requiere de un rango dinámico extenso de medición, sino también necesita manejar un rango de frecuencia de acuerdo al sistema de medición desarrollado.

3.1.1.1 Sensor de Corriente ALLEGRO ACS712 de 30A y 20A.

El sensor ACS712 de Allegro, es un sensor de corriente por efecto hall, que provee una solución económicas y precisas para el censado de corriente AC y DC, ya sea en el ambiente industrial o comercial. Este sensor funciona transformando un campo magnético surgido del paso de corriente por un alambre de cobre interno en el sensor, y convirtiendo este campo en un voltaje variable.

Esto significa que a mayor cantidad de corriente que tengamos, mayor voltaje vamos a tener en un pin. Este sensor viene en 3 modelos diferentes pero nosotros utilizaremos solo dos, el ACS712ELCTR-20A-T y el ACS712ELCTR-30A-T. La diferencia entre cada uno de los modelos es que las variaciones de voltaje en su pin de salida es siempre la misma, por ende para cualquier modelo su salida analógica varara entre 0 y 5v dándonos una mejor precisión en el modelo de 5A que en el de 30A.

El paquete del dispositivo permite una fácil implementación por el usuario. Las aplicaciones típicas incluyen el control de motor, detección de carga y administración, el modo de SWITCHES de las fuentes de poder, y protección por fallas de sobre corriente.

La salida del dispositivo cuenta con bucle (>Viot(q)).

Cuando una corriente creciente fluye a través de la pista de conducción primaria (desde los pines 1 y 2 a los pines 3 y 4), que es la pista utilizada para censar la corriente. La resistencia interna de esta pista conductora es de 1.2mohm típicamente, otorgando una baja perdida de voltaje.

El ajuste del conductor de cobre permite que el periodo de vida del dispositivo se extienda a condiciones de 5 x sobre corriente. Las terminales de la pista de conducción están eléctricamente aisladas de las carga del sensor (pin 5 a través del 8) esto permite que el sensor de corriente pueda ser usado en aplicaciones que requieran aislamiento eléctrico sin el uso de opto aisladores u otras técnicas de aislamiento costosas.

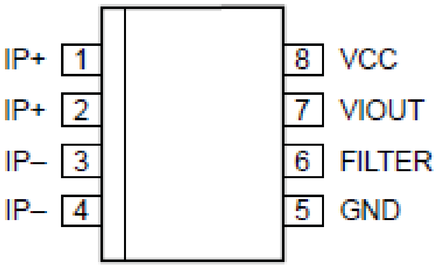


Figura 3.1 Sensor de corriente dibujo esquemático de pines.

3.3.1.2 Definición de las características de respuesta dinámica.

Retardo de propagación (t_{prop}) el tiempo requerido por el sensor para reflejar un cambio en la señal de corriente primaria. El retardo de propagación es atribuido a la carga inductiva sin el empaquetado IC linear, así como el bucle inductivo formado por el conductor geométrico primario. El retardo de propagación puede ser considerado como el tiempo compuesto de offset y puede ser compensado.

Tiempo de Respuesta ($I_{response}$) el intervalo de tiempo entre:

a) Cuando la señal de corriente primara alcanza el 90% de su valor final y
b) Cuando el sensor alcanza el 90% de su salida correspondiente a la corriente aplicada.

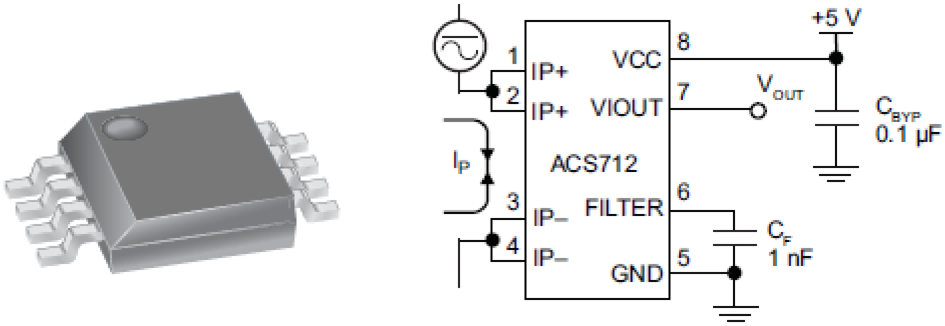


Figura 3.2 Diagrama de funcionamiento interno del sensor Allegro.

Tiempo de crecida o de aumento (Tt) el intervalo de tiempo entre:

a) Cuando el sensor alcanza el 10% del valor de su escala máxima y

b) Cuando alcanza el 90% del valor de su escala completa. El tiempo de incremento a un paso de respuesta es usado para derivar el ancho de banda de la corriente del sensor, en el que la integral de -3db-0.35/Tt ambas Tt y Tresponse son dramáticamente afectadas por las pérdidas de corriente observadas en los pines de tierra conductivos del IC.

3.1.1.3 Características técnicas.

Número de piezas	IP(A)	Sensibilidad (Typ) (mV/A)
ACS712ELCTR-20A-T	±20	100
ACS712ELCTR-30A-T	±30	66

Sensor de 20A | **Sensor de 30A**

Figura 3.3 ilustraciones de los sensores de 20 y 30 a

Especificaciones del sensor de 20A y 30A.

- Versión x05B (20 Amp)
- Bajo ruido
- El ancho de banda del dispositivo es seleccionado vía el pin FILTER
- Tiempo de respuesta de 5us ante cambios en la corriente.
- Ancho de banda de 80kHz
- Error de salida de 1.5% a 25°C
- 1.2mOhm de resistencia interna.

- 2.1 kVRMS de aislamiento de voltaje entre los pines 1-4 y los pines 5-8
- 5.0 VDC de alimentación.
- Sensibilidad de 66 a 185 mV/A
- Salida de voltaje proporcional a la corriente censada (AC o DC)
- Extremadamente estable.
- Histéresis magnética casi nula.
- Salida radiométrica

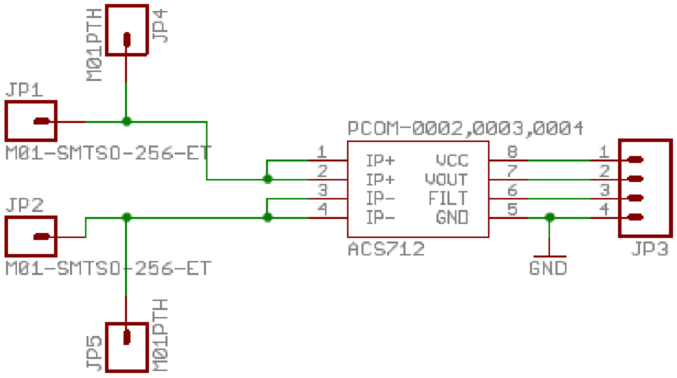


Figura 3.4 Diagrama esquemático del sensor de 20a y 30a

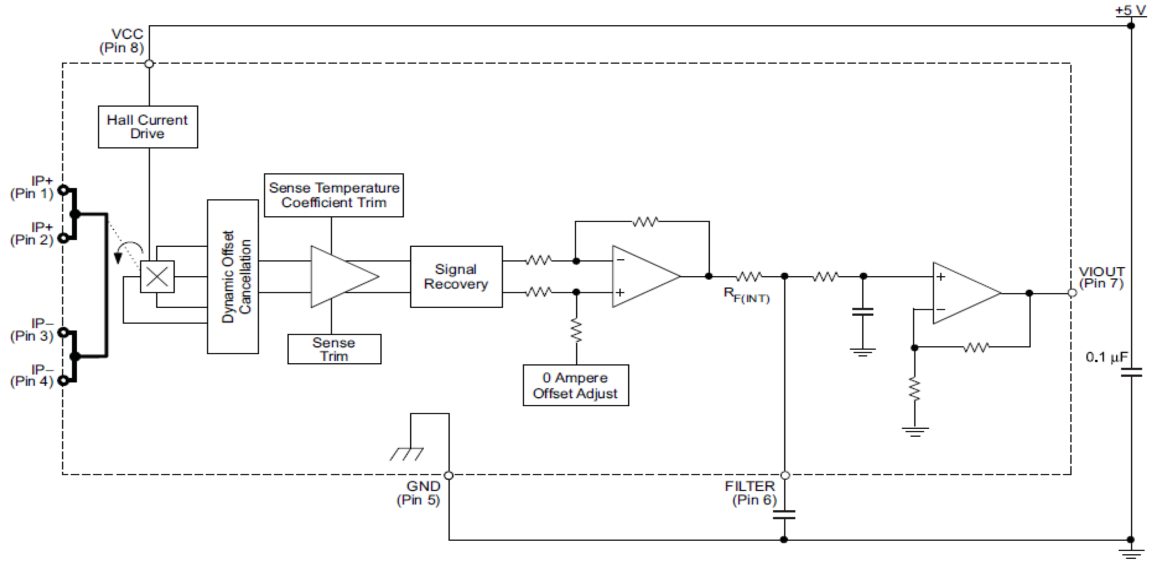


Figura 3.5 Diagrama de bloque funcional del sensor de 20a y 30a

3.1.1.4 conexiones

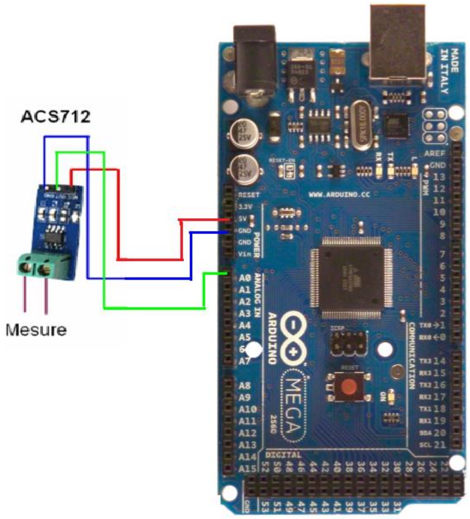


Figura 3.6 Diagrama de conexión Shield Allegro/Arduino

3.1.1.5 LTC1966FB Sensor Convertidor RMS-DC

Es un convertidor que utiliza una técnica computacional novedosa patentada delta sigma. La circuitería interna delta sigma del ltc1966 lo hace simple de usar, más preciso, menos consumo de poder y dramáticamente más flexible que los convertidores rms-dc convencionales.

El ltc1966 acepta entradas de una sola terminación o señales diferenciales. A diferencia de los anteriores rms-dc convertidores disponibles, la linealidad superior del ltc permite una calibración libre a cualquier entrada de voltaje.

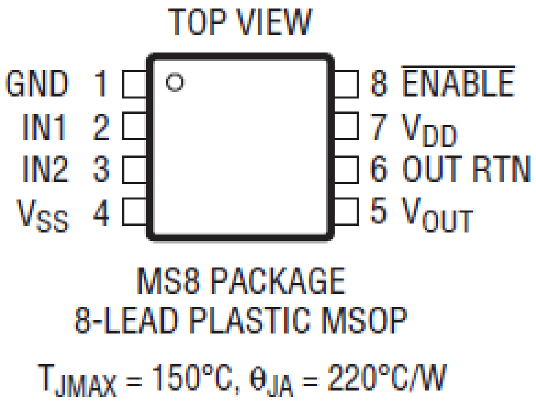


Figura 3.7 Convertidor RMS-DC

V_{DD} to GND	-0.3V to 7V
V_{DD} to V_{SS}	-0.3V to 12V
Input Currents (Note 2)	± 10Ma
V_{SS} to GND	-7V to 0.3V
Output Current (Note 3)	± 10Ma
ENABLE Voltage	V_{SS} -0.3V to V_{SS} + 12V
OUT RTN Voltage	V_{SS} -0.3V to V_{DD}
Operating Temperature Range (Note 4)	
LTC1966C/LTC1966I	-40°C to 85°C
LTC1966H	-40°C to 125°C
LTC1966MP	-55°C to 125°C
Specified Temperature Range (Note 5)	
LTC1966C/LTC1966I	-40°C to 85°C
LTC1966H	-40°C to 125°C
LTC1966MP	-55°C to 125°C
Maximum Junction Temperature	150°C
Storage Temperature Range	65°C to 150°C
Lead Temperature (Soldering, 10 Sec)	300°C

3.1.1.5.1 Funcionamiento de los pines.

Funciones de los pines

Gnd (pin1): tierra

IN1 (PIN2): entrada diferencial. Acoplamiento DC, (la polaridad es irrelevante)

IN2 (PIN3) : entrada diferencial, acoplamiento DC(la polaridad es irrelevante)

Vss(PIN4): alimentación de voltaje negativa. GND a -5-5V.

Vout (PIN5): salida de voltaje. Esta es de alta impedancia. El promedio RMS se cumple con la implementación de un capacitor de este nodo al OUT RTN. La función de transferencia está dada por:

(Vout – OUT RTN) = sqr(Average(IN2-IN1)¨2)

OUT RTN (PIN 6): retorno de salida. La salida de voltaje es creada relativamente a este pin. Los pines Vout y OUT RTN no está balanceados y este pin debe ser ajustado a una baja impedancia, en ambas AC y DC. Además es físicamente enlazado a GND, pero puede ser enlazado a cualquier voltaje arbitrario, Vss<OUT RTN<(VDD- Max Output). Los mejores resultados son obtenidos cuando OUT RTN=GND.

VDD (PIN7): alimentación de voltaje positiva. 2.7V a 5.5V.

ENABLE (PIN 8): para una operación normal ponga a tierra o a un cero lógico incluso a Vss.

Definición RMS.

La amplitud rms es la constante, aplicable y estándar manera de medir y de comparar señales dinámicas de todos los tamaños y formas, en estado simple, la amplitud rms es el potencial de calentamiento de una señal de onda dinámica. 1vrms de onda AC generara el mismo calor en una carga resistiva así como 1vdc. Matemáticamente el voltaje rms es el resultado de la raíz cuadrada del voltaje elevado al cuadrado.

Alternativas del RMS.

Otras formas de cuantificar las ondas dinámicas incluyen la detección de picos de voltaje y su rectificación promedio, en ambos casos un resultado promedio de dc, pero el valor solo es acerado al elegir el tipo de onda al cual esta calibrado, típicamente son ondas senoidales.

3.1.1.6 Como Opera un Convertidor RMS-CD

Los convertidores monolíticos RMS-DC utilizan una computación implícita para calcular el valor rms de una señal de entrada.

La construcción fundamental de este bloque es un divisor/multiplicador analógico, como se muestra en la figura. El análisis de esta topología es sencillo y comienza al identificar las entradas y las salidas del filtro pasa bajos. La entrada del filtro pasa bajos es el cálculo resúltale del multiplicador/divisor, el filtro pasa bajos tomara el promedio de esto para generar la salida.

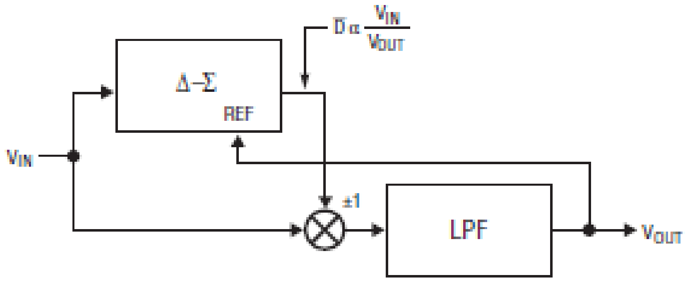


Figura 3.8 Topología del LTC1966.

3.1.1.7 Como opera el LTC1966.

El Ltc1966 es una topología nueva para a conversión de rms a dc en el que delta sigma actúan como modulador del divisor, y un simple SWITCH de polaridad es utilizado como el multiplicador.

El modulador delta sigma contiene una salida de un solo bit que promedia el ciclo de trabajo y será proporcional al rango de la señal de entrada dividida con la salida, el delta sigma es un modulador de segundo orden con una linealidad excelente, la salía de n solo bit es utilizada selectivamente para combinar o invertir la señal de entrada, una vez más este es n circuito con una linealidad excelente, porque pera solamente a dos puntos a ganancia +-1, la multiplicación promedio efectiva será en la línea recta entre estos dos puntos, la combinación de esos dos elementos nuevamente tendrán un filtro pasa bajos en la señal de entrada proporcional que, como se muestra arriba resulta en la conversión rms-dc.

El filtro pasa bajos realiza el promedio de la función rms y debe ser una esquina de baja frecuencia que la más baja frecuencia de interés, para mediciones de frecuencia linear el filtro es simplemente muy grande para ser implementado en un chip, pero el ltc solo necesita un capacitor en la salida para implementar el filtro pasa bajos, el usuario puede seleccionar este capacitor dependiendo del rango de frecuencia y los requerimientos de ajuste de tiempo.

3.1.2 Linealidad de un convertidor RMS DC

La linealidad puede ser como una propiedad básica para un dispositivo que implementa una función que incluye dos procesos no lineales. Como sea un convertidor rms-dc tiene una función de transferencia rms volts a dc volts de salida, que idealmente deben tener una función de transferencia 1:1.

Una vista más completa a la linealidad utiliza el modelo más simple mostrado abajo, ahí un núcleo ideal rms que está corrompido por ambas circuiterías de entrada y de salida que contienen funciones de transferencia incorrectas, como se puede notar el offset de entrada es introducido en la circuitería de entrada, mientras el offset de salida esta introducido en la circuitería de salida.

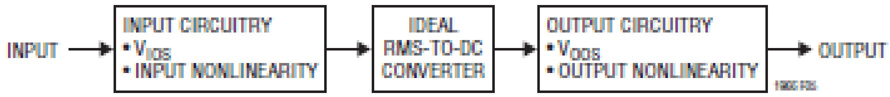


Figura 3.9 Linealidad de un convertidor RMS DC

Cualquier no linealidad que ocurra en la circuitería de salida corrompe la salida de la función de transferencia RMS-DC, una no linealidad en la función de entrada típicamente corromperán la función de transferencia aun más, simplemente porque una entrada AC, la conversión RMS a DC tendrá una o linealidad promedio de todo un rango de valores de entrada juntos.

Pero la entrada no lineal aun causara problemas en un convertidor RMS-DC porque corromperá la precisión conforme la forma de la señal de entrada cambie, así mismo un convertidor rms-dc convertirá cualquier onda de entrada a una salida dc, la precisión no es necesaria para todas las formas de onda.

3.1.3 Arduino

El Arduino es una plataforma de desarrollo de código abierto, basada en una tarjeta con entradas y salidas, y un entorno de programación que implementa el lenguaje Processing/Wiring. Arduino puede ser utilizado para desarrollar objetos autónomos interactivos o puede ser conectado con el software de la computadora (por ejemplo, Flash, Processing, MaxMSP). El IDE de código abierto puede ser descargado de forma gratuita para Mac OS X, Windows y Linux.

El Arduino Mega es una tarjeta de desarrollo basada en el ATmega2560. Tiene 54 pines de entrada/salida digital (de los cuales 14 pueden ser usados como salidas de PWM), 16 entradas analógicas, 4 UARTs (puertos seriales), un oscilador de cristal de 16 *MHz*, una conexión USB, un conector de alimentación, un conector para ICSP, y un botón de reinicio. Contiene todo lo necesario para usar el microcontrolador, basta con conectarlo a un ordenador con un cable USB o energizarlo con una fuente de poder o batería para comenzar a usarlo.

El Arduino Mega es compatible con la mayoría de los shields diseñados para el Arduino Duemilanove o Diecimila.

El Arduino Mega 2560 R3 también añade pines SDA y SCL al lado del pin AREF. Además, hay dos nuevos pines colocados cerca del pin de RESET. Uno de ellos es el IOREF que permiten que los shields se adapten al voltaje suministrado desde la tarjeta. El otro es uno que no está conectado reservado para usos futuros. El arduino Mega 2560 R3 trabaja con todos los shields existentes y se puede adaptar a nuevos shields que utilizan estos pines adicionales.

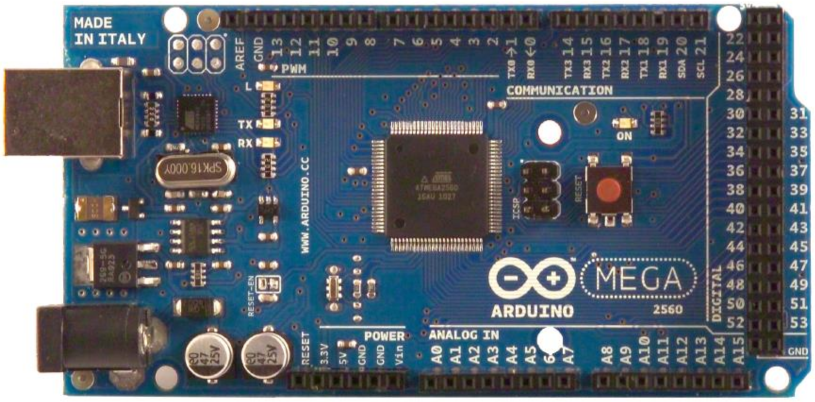
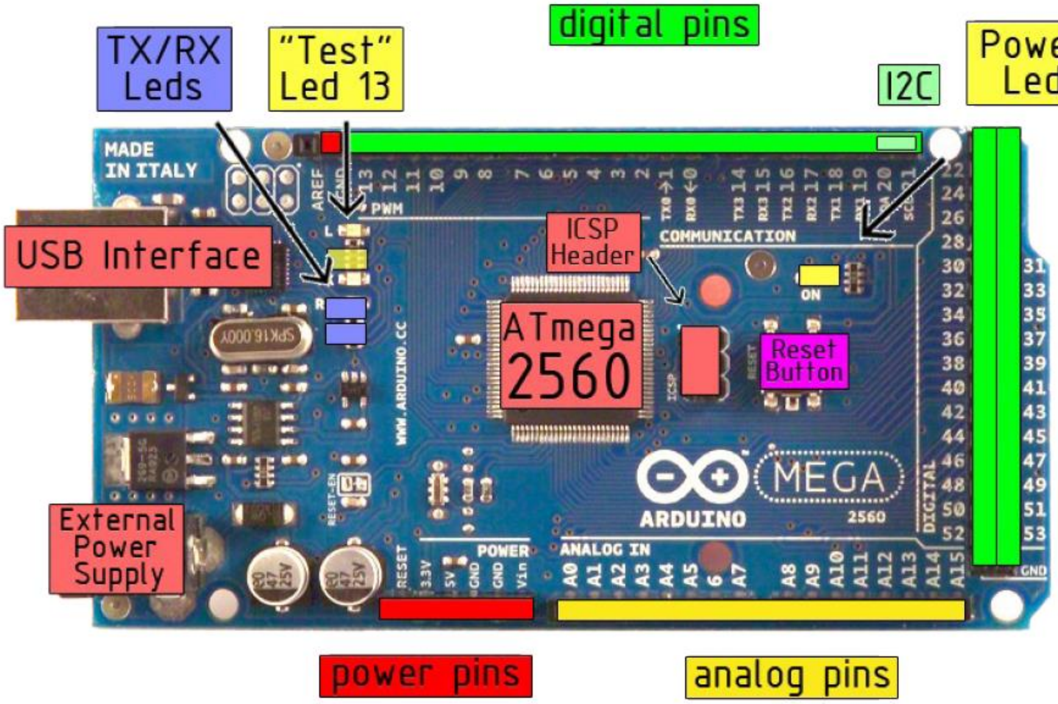


Figura 3.10 Arduino Mega

3.1.3.1 Características:

- Microcontrolador ATmega2560
- Operación del voltaje 5V
- Voltaje de entrada 7-12 *V*
- 54 pines de entrada-salida digital (14 de los cuales se pueden usar como salida de PWM)
- 16 puertos analógicos
- Memoria Flash de 256 *KB*
- SRAM 8KB
- EEPROM 4KB



3.11 Descripción de puertos y funciones Arduino Mega

Para entender mejor el comportamiento del Arduino hay que saber primero que es un Arduino. El Arduino es una plataforma de prototipos electrónica de código abierto (open-source) basada en hardware y software flexibles y fáciles de usar. Está pensado para artistas, diseñadores, como hobby y para cualquiera interesado en crear objetos o entornos interactivos.

El microcontrolador de la placa se programa usando el <<Arduino Programming Language>> (basadoen Wiring1) y el <<Arduino Development Environment>> (basado en Processing2). Los proyectos de Arduino pueden ser autónomos o se pueden comunicar con software en ejecución en un ordenador (por ejemplo con Flash, Processing, MaxMSP, etc.).

Las placas se pueden ensamblar a mano o encargarlas preensambladas; el software se puede descargar gratuitamente. Los diseños de referencia del hardware (archivos CAD) están disponibles bajo licencia open-source, por lo que eres libre de adaptarlas a tus necesidades. Arduino recibió una mención honorífica en la sección Digital Communities del Ars electrónica Prix en 2006.

Hay muchos otros microcontroladores y plataformas microcontroladoras disponibles para computación física. Parallax Basic Stamp, Netmedia's BX-24, Phidgets, MIT's Handyboard, y muchas otras ofertas de funcionalidad similar. Todas estas herramientas toman los desordenados detalles de la programación de microcontrolador y la encierran en un paquete fácil de usar. Arduino también simplifica el proceso de trabajo con microcontroladores, pero ofrece algunas ventajas para profesores, estudiantes interesados sobre otros sistemas como:

- Barato: Las placas Arduino son relativamente baratas comparadas con otras plataformas microcontroladoras. La versión menos cara del modulo Arduino puede ser ensamblada a mano, e incluso los módulos de Arduino preensamblados cuestan menos de $50.

- Multiplataforma: El software de Arduino se ejecuta en sistemas operativos Windows, Macintosh OSX y GNU/Linux. La mayoría de los sistemas microcontroladores están limitados a Windows.

- Entorno de programación simple y claro: El entorno de programación de Arduino es fácil de usar para principiantes, pero suficientemente flexible para que usuarios avanzados puedan aprovecharlo también. Para profesores, está convenientemente basado en el entorno de programación Processing, de manera que estudiantes aprendiendo a programar en ese entorno estarán familiarizados con el aspecto y la imagen de Arduino.

- Código abierto y software extensible: El software Arduino está publicado como herramientas de código abierto, disponible para extensión por programadores experimentados. El lenguaje puede ser expandido mediante librerías C++, y la gente que quiera entender los detalles técnicos pueden hacer el salto desde Arduino a la programación en lenguaje AVR C en el cual está basado. De forma similar, puedes añadir código AVR-C directamente en tus programas Arduino si quieres.

- Código abierto y hardware extensible: El Arduino está basado en microcontroladores ATMEGA8 y ATMEGA168 de Atmel. Los planos para los módulos están publicados bajo licencia Creative Commons, por lo que diseñadores experimentados de circuitos pueden hacer su propia versión del módulo, extendiéndolo y mejorándolo. Incluso usuarios relativamente inexpertos pueden construir la versión de la placa del módulo para entender cómo funciona y ahorrar dinero.

3.1.3.2 Hardware.

Hay múltiples versiones de la placa Arduino. La mayoría usan el ATmega168 de Atmel, mientras que las placas más antiguas usan el ATmega8.

Placas E/S

- Diecimila: Esta es la placa Arduino más popular. Se conecta al ordenador con un cable estándar USB y contiene todo lo que necesitas para programar y usar la placa. Puede ser ampliada con variedad de dispositivos: placas hijas con características específicas.
- Nano: Una placa compacta diseñada para uso como tabla de pruebas, el Nano se conecta al ordenador usando un cable USB Mini-B.
- Bluetooth: El Arduino BT contiene un modulo bluetooth que permite comunicación y programación sin cables. Es compatible con los dispositivos Arduino.
- LilyPad: Diseñada para <<aplicaciones listas para llevar>>, esta placa puede ser conectada en fábrica, y un estilo sublime.
- Mini: Esta es la placa más pequeña de Arduino. Trabaja bien en tabla de pruebas o para aplicaciones en las que prima el espacio. Se conecta al ordenador usando el cable Mini USB.
- Serial: Es una placa básica que usa RS232 como un interfaz con el ordenador para programación y comunicación. Esta placa es fácil de ensamblar incluso como ejercicio de aprendizaje.
- Serial Single Sided: Esta placa está diseñada para ser grabada y ensamblada a mano. Es ligeramente más grande que la Diecimila, pero aun compatible con los dispositivos.

3.1.3.3 Comunicación.

El Arduino Diecimila tiene un número de infraestructuras para comunicarse con un ordenador, otro Arduino, u otros microcontroladores. El ATmega168 provee comunicación serie UART TTL (5 V), la cual está disponible en los pines digitales 0 (Rx) y 1 (Tx). Un FTDI FT232RL en la placa canaliza esta comunicación serie al USB y los drivers FTDI (incluidos con el software Arduino) proporcionan un puerto de comunicación virtual al software del ordenador. El software Arduino incluye un monitor serie que permite a datos de texto simple ser enviados a y desde la placa Arduino.

Una librería SoftwareSerial 7 permite comunicación serie en cualquiera de los pines digitales del Diecimila.

El ATmega168 también soporta comunicación 12C (TWI) y SPI. El software Arduino incluye una librería Wire para simplificar el uso del bus 12C8. Para usar la comunicación SPI, consultar el esquema del ATmega168.

3.1.3.4 Programación.

El Arduino Diecimila puede ser programado con el software Arduino. El ATmega168 del Arduino Diecimila viene con un bootloader pregrabado que te permite subirle nuevo código sin usar un programador hardware externo. Se comunica usando el protocolo original STK500.

También puedes saltar el bootloader y programar el ATmega168 a través de la cabecera ICSP (In-Circuit Serial Programming).

Reseteo Automático (Software)

En lugar de requerir una pulsación física del botón de reset antes de una subida, el Arduino Diecimila está diseñado de forma que permite ser reseteado por software en ejecución en una computadora conectada. Una de las líneas de control de flujo de hardware (DTR) del FT232RL está conectada a la línea de reset del ATmega168 a través de un condensador de 100 nF. Cuando esta línea toma el valor LOW, la línea reset se mantiene el tiempo suficiente para resetear el chip. La versión 0009 del software Arduino usa esta capacidad para permitirte cargar código simplemente presionando el botón upload en el entorno Arduino.

Esto significa que el bootloader puede tener un tiempo de espera más corto, mientras la bajada del DTR puede ser coordinada correctamente con el comienzo de la subida.

3.1.3.5 Obtener una placa ARDUINO y un cable.

El Arduino Mega es una placa que contiene todo lo que necesitas para empezar a trabajar con electrónica y programación de microcontrolador. También necesitas un cable USB estándar (del tipo que conectarías a una impresora USB, por ejemplo).

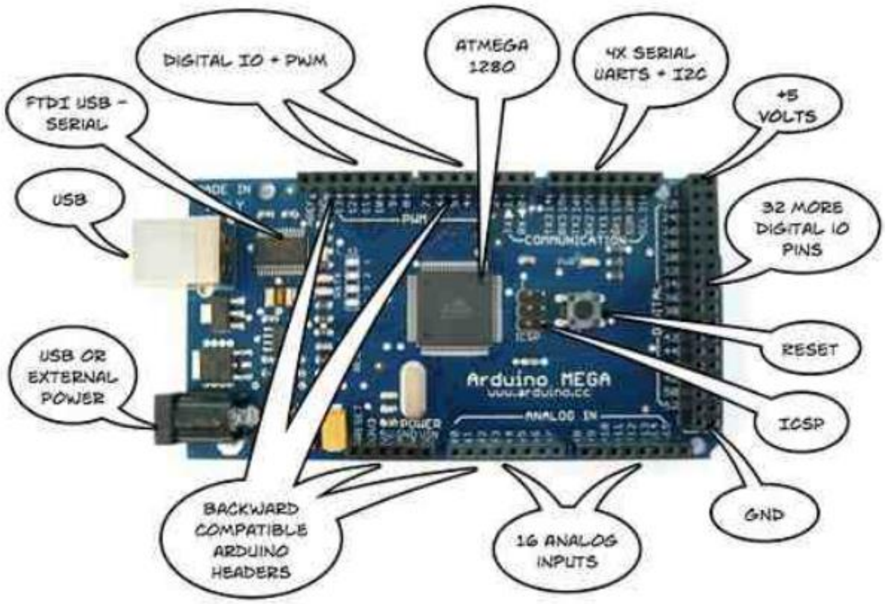


Figura 3.12. Descripción de componentes de la placa ARDUINO MEGA.

3.1.4 Sensor de Voltaje Pico

La principal limitación de los diodos de silicio es que no son capaces de rectificar voltajes por debajo de 0.6v. Siempre que haya voltajes de entrada positivos de producirá un voltaje de salida, aun cuando dichos voltajes estén por debajo de 0.6v . Los circuitos rectificadores de media onda transmiten solamente la mitad de un ciclo de una señal y elimina el otro, al limitar su salida a cero volts. La mitad del ciclo que si se transmite puede estar invertida o no.

Además de servir para rectificar de manera precisa una señal, diodos y amplificadores operacionales se conectan para construir circuitos detectores de pico. Este tipo de circuitos sigue los picos de voltaje de una señal y almacena en un capacitor el valor máximo que se haya alcanzado (durante un tiempo casi indefinido). Cuando llega una señal pico mayor, se almacena este nuevo valor. El voltaje de pico más elevado se almacena hasta que se produce la descarga del capacitor por medio de un interruptor mecánico o electrónico. A este circuito detector de pico también se le conoce como circuito seguidor o retenedor o seguidor de pico. Vemos también que al invertir los diodos en este circuito se obtiene en vez de un seguidor de pico, un seguidor de valle.

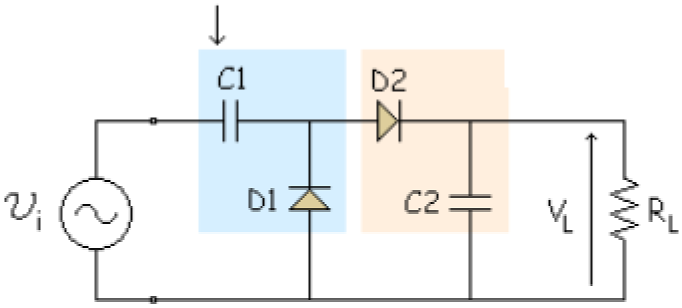


Figura 3.13 Diagrama ilustrativo del funcionamiento de un sensor de voltaje pico

3.2.-Desarrollo del proyecto

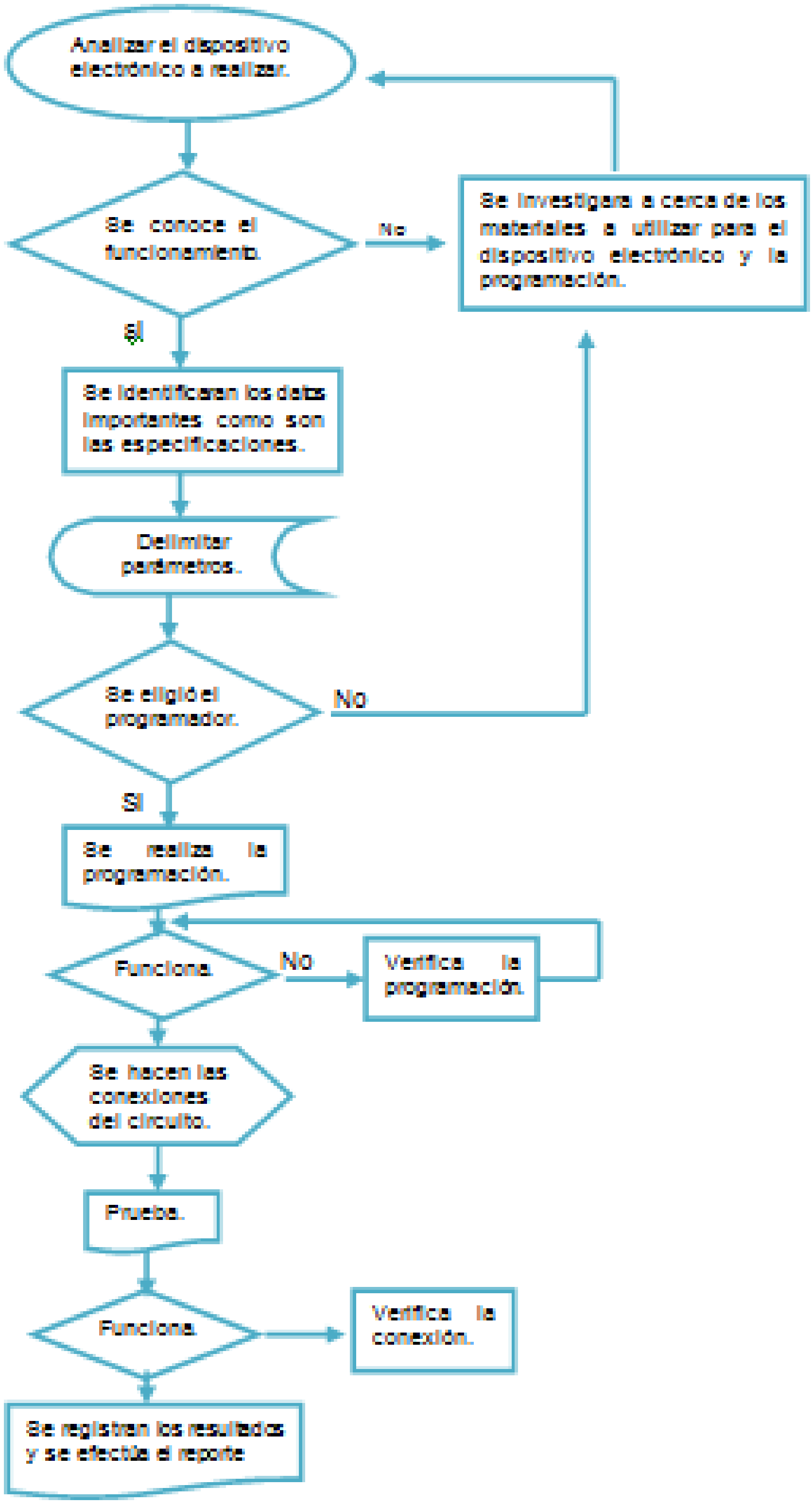
3.2.1 Descripción del sistema.

El sistema realizado de instrumentación y protección del restaurador dinámico de voltaje está formado por dos partes principales, hardware y software. Donde la parte del software contiene la parte de programación del Arduino que sirve para que se pueda registrar y desplegar los datos en tiempo real en forma de graficas en la computadora. El hardware es la parte donde podemos encontrar el circuito que nos representa el control para el monitoreo y el cálculo de energía, en el cual se llevan a cabo los cálculos de medición, esto para que junto al software se esté verificando que la energía de entrada no se dispare y si este llegara a pasar, el sistema se encargara de bloquear la carga de entrada para que no dañe el dispositivo que se maneje.

3.2.2 Procedimiento y descripción de las actividades realizadas.

En esta etapa del proyecto después de haber dado a conocer las bases teóricas del Regulador Dinámico de Voltaje y las bases teóricas necesarias para la comprensión del sistema de regulación de voltaje e instrumentación del mismo que es el objetivo general a lograr, se darán a conocer la metodología donde podremos saber específicamente que es lo que se realizara para poder cumplir con la finalidad del proyecto. Describiendo cada pasó de la manera más breve en que fue realizada hasta concluirla.

3.2.3 Metodología



3.2.4 Pasos de cómo se llevo a cabo el prototipo.

1. Para comenzar a trabajar con el proyecto lo primero que se realizo fue un análisis para saber si, al meter un sensores de corrientes con una entrada alta como 20 o 30 amperes, se lograra reducir el voltaje q circulara ya que estaría pasando un voltaje de corriente alterna la cual queremos pasar a un nivel de tensión inferior por lo cual usamos un transformador de 120 A a 12V para poder llevar a cabo dicha acción. Después de bajarlo a un volate considerado lo que se proseguiría es meter dicha señal de entrada al arduino para que este mediante una programación que se ejecutara en una PC se pueda monitorear la entrada de voltaje q está recibiendo el Arduino.

   Pero no solo eso, también por medio de un sensor de Vrms checar el voltaje que estábamos recibiendo de entrada y del integrado en Vrms para poder comparar las diferencias al igual que el voltaje que nos proporcionaba pico a pico, todo esto con la finalidad de poder saber cómo estaba funcionado el voltaje que estábamos recibiendo en parámetros distinto con un determinado valor de Amperaje. Todo esto se llevaría a cabo con la finalidad de que al momento de que algo fallara en el sistema y entrara un voltaje total de corriente alterna el programa pudiera detectarlo y bloquear el circuito abriéndolo para que la energía que circula no pueda pasar y perjudicar el sistema que se trabaje.

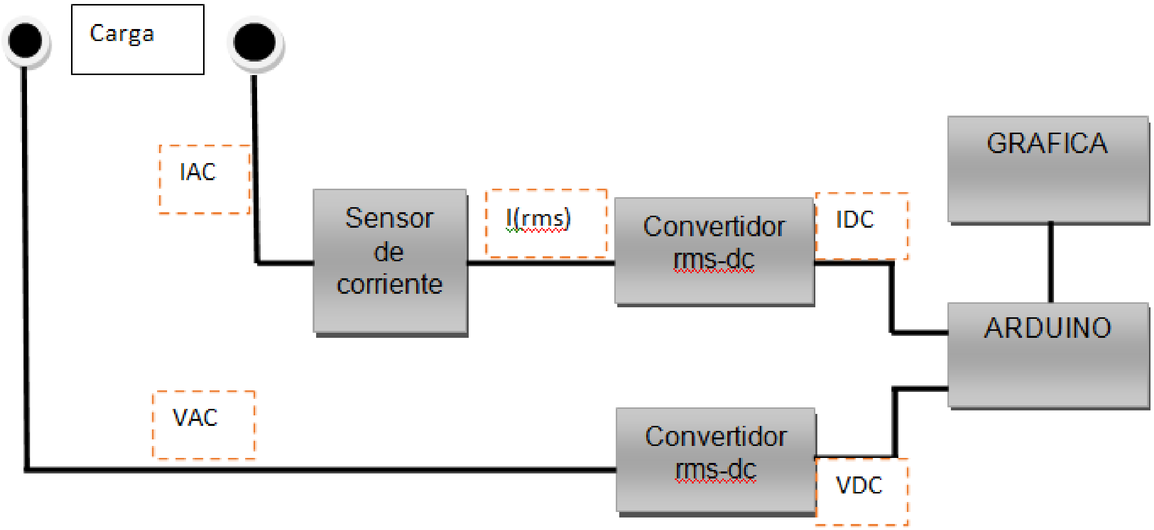


Figura 3.14 Diseño de bloques del prototipo

2. Desde que se empezó hacer un bosquejo del cto. Que se quería hacer se eligió el microcontrolador con el que se quería trabajar, que fue el Arduino mega, y se decidió que con base a este se haría la programación para la protección.

3. Se empezó con las investigaciones sobre los tipos de sensores que podrían servirnos para nuestro proyecto, se analizaron las características de ellos y se opto por decidir que se trabajaría con sensores de corriente para el suministro de entrada, un integrado para que nos pudiera proporcionar el Vrms a Dc y otra aplicación que nos permitiera ver el voltaje pico a pico que manejaba. Todo esto con la finalidad que se pudiera ver el comportamiento de la corriente.

4. Una vez elegido los sensores que se utilizaron para el dispositivo electrónico, se empezaron a valorar las características especificas de cada sensor ya que con estos parámetros importantes lograríamos elegir una delimitación de el rango que se quiere manejar para proteger el sistema, ya que gracias al conocer la importancia de cada sensor, su funcionamientos específicos, solo de esa manera se podría delimitar esa parte importante y fundamental para el dispositivo.

5. Cuando se obtuvieron las características especificas de los sensores se prosiguió a la elaboración de la programación como se observara mas adelante y con su explicación en cada parte de ella.

6. Al finalizar con el programa se llevo a cabo las pruebas para corroborar con el funcionamiento que debería hacer el sensor esto con la finalidad que pudiéramos saber el comportamiento de la corriente que se estaba suministrando y la variación que hacía en cada sensor para el voltaje.

7. Una vez corroborado que cada sensor ejecutaba su función correctamente se hizo la implementación para checar que el funcionamiento conjunto funcionara de la misma manera sin ningún problema, se tuvieron complicaciones pero al final se pudo obtener los resultados finales que se quería, que fue el monitoreo de las señales que entraban como las de salida.

8. Con la implementación del software se pudo monitorear mediante las graficas las señales diferentes con las que se hicieron las pruebas. El circuito que se observa a continuación es parte del prototipo.

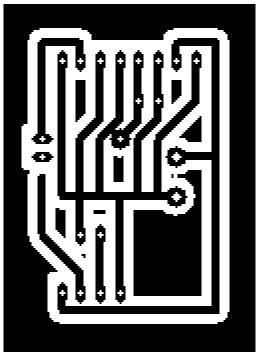


Figura 3.15 Pista del Prototipo del sensor

Para poder elaborar este dispositivo electrónico y la programación para la protección se estimo el tiempo de 16 semanas, dividiendo el tiempo en pequeñas fracciones para poder trabajar con cada parte del dispositivo y que se le diera la dedicación necesaria para que el objetivo se cumpliera.

Actividades	Semanas															
	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16
Investigaciones sobre el estado de medición y protección en equipos DVR.	■	■	■	■	■	■										
Seleccionar los sensores para la protección del DVR y el tpo de programación que se realizara .				■	■	■	■									
Diseño y construcción del dispositivo y el acondicionamiento de los sensores seleccionados para la realización de la programación								■	■	■	■	■	■			
Prueba y medición de los sensores por separado y registro de resultados													■	■		
Interconexión de la instrumentación electrónica y prueba de ella.														■	■	■

3.16 Tiempo estimado para las actividades

Mientras se trabajo con el diseños del prototipo hasta que se finalizo, el trabajo estuvo supervisado por mi asesor el Ing. José Ángel Zepeda Hernández y se trabajo en el laboratorio de ingeniería electrónica. Los materiales utilizados fueron proporcionados por mi asesor.

3.2.5 Materiales y dispositivos electrónicos

- 1 Arduino Mega 2560
- 1 Sensor de corriente ACS712ELCTR-30A-T
- 1 Sensor de corriente ACS712ELCTR-20A-T
- 1 Sensor de voltaje pico
- 1 sensor de voltaje Vrms LTC1966CM58
- COMPUTADORA
- PROTOBOARS.

CAPITULO IV

4.1 Programa del prototipo

Después de haber conocido los elementos que se utilizaran para generar este prototipo, así como la instalación de la plataforma Arduino y las características principales que se utilizaron como el puerto serial, se procedió a crear un código utilizando la plataforma para enlazarla con C#, donde se genero el software que va a mostrar los datos censados por el Arduino, en este caso y como se explico en las limitaciones del proyecto solo se incluyo 1 sensor al Arduino, ya que por falta de los sensores para su parametrización, así como de tiempo por investigaciones se opto por la representación del mismo con un potenciómetro.

```
void setup()

{
  Serial.begin(115200);
  Inicializa();
}
```

En esta sección del código se configura el programa y se inicializa el puerto serial, en lugar de 9200 baudios se configura a 115200 para que la velocidad de muestreo sea mucho mayor, y se pasa a la función de inicialización de parámetros.

```
void loop()
{
  Get_Instruccion();
}
```

En esta sección del código llamamos al bucle e invocamos a la función donde hace referencia a las instrucciones de lectura y escritura de datos.

```
int Get_Instruccion()
{
  if(Serial.available())
  {
      switch(Serial.read())
      {
         case 'r':
           Analog_Read();
```

```
        break;

        case 's':
          Salir();
        break;
     }
  }
}

void Salir()
{
  Inicializa ();
}
```

En esta sección del programa, de acuerdo a los parámetros que se ejecute en ese momento se lleva a cabo un menú, por así decirlo, de opciones internas, o una comparación donde el programa detecta la disponibilidad del puerto serial, para tomar una acción con respecto a lo que interprete.

```
void Analog_Read()
{
    int resultado = 0;
    resultado = analogRead(0);
    Serial.write(resultado / 4);
}
```

En esta sección es donde el programa lee lo que se está recibiendo por el puerto serial configurado y conectado en el pin 0 del Arduino y lo transmite en bits para ser interpretado por el compilador asociado.

```
void Inicializa()
{
  int caracter[6];
  int i = 0;
  caracter[0] = 's';
  caracter[1] = 'q';
  caracter[2] = 'u';
  caracter[3] = 'a';
  caracter[4] = 'r';
  caracter[5] = 'e';

  while(1)
  {
```

```
     if(Serial.available())
    {
       if(Serial.read() == caracter[i])
       {
         if(i == 5)
         {
           Serial.write('a');
           break;
         }
         i++;
       }
     }
  }

}
```

En esta última sección no quiere decir que sea la última en leerse, es la función de inicialización invocada al principio del programa a la cual se estableció un menú implícito, para facilitar los parámetros de decisiones del Arduino con respecto al compilador y ver las propiedades de lectura del puerto así como su disponibilidad.

PROGRAMA EN C#

El programa visual se desarrolla en C# ya que es un software de programación de alto nivel y muy versátil sobre todo con el Arduino, el programa en si es extenso ya que se utilizo la parte visual y la configuración de cada elemento.

Programa principal para establecer comunicación con el Arduino.

```
using System;
using System.Collections.Generic;
using System.Linq;
using System.Text;
using System.IO.Ports;
using System.Windows.Forms;

namespace Practica2_LabDSM
{
    class Arduino
    {
        SerialPort serialPort;
        bool ok;

        public Arduino(string portname, int bauds)
        {
```

```
            try
            {
                serialPort = new SerialPort(portname, bauds);
                ok = true;
            }
            catch (Exception ex)
            {
                MessageBox.Show("Error en la asignacion del puerto del arduino. " + ex,
"Error de Inicializacion de Puerto", MessageBoxButtons.OK,
MessageBoxIcon.Error);
                ok = false;
            }
        }

        public bool Inicializa_Comunicacion()
        {
            if (ok)
            {
                try
                {
                    if (!serialPort.IsOpen) //Si el puerto esta cerrado
                        serialPort.Open();
                    serialPort.ReadTimeout = 95;
                    serialPort.Write("square");
                    if (serialPort.ReadChar() == 'a')
                    {
                        MessageBox.Show("Comunicacion Establecida", "Correcto",
MessageBoxButtons.OK, MessageBoxIcon.Information);
                    }
                    return true;
                }
                catch (Exception ex)
                {
                    MessageBox.Show("Error en la asignacion del puerto del arduino. " +
ex, "Error de Inicializacion de Puerto", MessageBoxButtons.OK,
MessageBoxIcon.Error);
                    return false;
                }
            }
            else
            {
                return false;
            }
        }

        public void CierraPuerto()
        {
```

```
            if (serialPort.IsOpen)
                serialPort.Close();
        }

        public void Stop()
        {
            if (ok)
            {
                serialPort.Write("s");
            }
        }


        public int GetLectura()
        {
            int lectura = -1;
            try
            {

                if (ok)
                {
                    serialPort.Write("r");
                    lectura = serialPort.ReadChar();
                }
            }
            catch
            {
                MessageBox.Show("Error Interno al obtener Lectura, la aplicacion se
cerrara", "Error", MessageBoxButtons.OK, MessageBoxIcon.Error);
                Application.Exit();
            }
            return lectura;
        }

    }
}
```

Programa de enlace entre el código principal y el Form donde se adquieren los datos y las configuraciones de comunicación con el Arduino, el software y el puerto serial, es la ejecución de los eventos en el Form.

```
using System;
using System.Collections.Generic;
using System.ComponentModel;
using System.Data;
```

```
using System.Drawing;
using System.Linq;
using System.Text;
using System.Windows.Forms;
using System.IO.Ports;

namespace Practica2_LabDSM
{
    public partial class Form_Main : Form
    {
        Arduino arduino = null;
        int[] muestreo;

        public Form_Main()
        {

            InitializeComponent();
            muestreo = new int[50];
        }

        private void groupBox_PuertoSerial_Enter(object sender, EventArgs e)
        {

        }

        private void Form_Main_Load(object sender, EventArgs e)
        {
            serial_combo(this.comboBox_Puertos);
            for (int i = 0; i < 50; i++)
            {
                //this.chart_Datos.Series[0].Points.AddY(i);
                muestreo[i] = 0;
            }

        }

        private void button_Iniciar_Click(object sender, EventArgs e)
        {
            if (this.comboBox_Puertos.Items.Count > 0)
            {
                if (this.comboBox_Puertos.SelectedIndex > -1)
                {//Si esta seleccionado algun puerto
                    arduino = new Arduino(this.comboBox_Puertos.Text,
115200);//Inicializamos arduino a 115200 bauds
                    arduino.Inicializa_Comunicacion();
                    this.button_Iniciar.Enabled = false;
                    this.button_Stop.Enabled = true;
```

```
            timer.Enabled = true;
            timer.Start();
        }
    }
}


//Metodo que primero limpia y despues llena la comboBox pasada como
//parametro con los puertos serie disponibles
private void serial_combo(ComboBox combo)
{
    combo.Items.Clear();
    string[] ports = SerialPort.GetPortNames();

    foreach (string port in ports)
    {
        combo.Items.Add(port);
    }
}

private void Form_Main_FormClosed(object sender, FormClosedEventArgs e)
{
    try
    {
        arduino.CierraPuerto();
    }
    catch
    {
        ;
    }
}

private void timer1_Tick(object sender, EventArgs e)
{
    Get_Lectura();
}

private void button_Stop_Click(object sender, EventArgs e)
{
    try
    {
        arduino.Stop();
        arduino.CierraPuerto();
        arduino = null;
        this.button_Stop.Enabled = false;
        this.button_Iniciar.Enabled = true;
        this.timer.Stop();
```

```
        this.timer.Enabled = false;
    }
    catch
    {
        arduino = null;
    }
}

private void Get_Lectura()
{
    int punto = 0;
    if (arduino != null)
    {
        punto = arduino.GetLectura();
        this.textBox_caracter.Text = Convert.ToChar(punto).ToString();
        ActualizaGrafica(punto);
    }
}

private void ActualizaGrafica(int punto)
{
    RecorreArray(punto);
    this.chart_Datos.Series[0].Points.Clear();
    for (int i = 0; i < 50; i++)
    {
        this.chart_Datos.Series[0].Points.AddY(muestreo[i]);
    }
}

private void RecorreArray(int punto)
{
    int[] muestreoAux = new int[50];
    for (int i = 0; i < 49; i++)
    {
        muestreoAux[i + 1] = muestreo[i];
    }
    muestreoAux[0] = punto;
    muestreo = muestreoAux;
}

private void button_Refresh_Click(object sender, EventArgs e)
{
    serial_combo(this.comboBox_Puertos);
}

private void chart_Datos_Click(object sender, EventArgs e)
{
```

```
        }

    }
}
```

Programa de diseno de la interfaz visual y su configuracion

```
namespace Practica2_LabDSM
{
    partial class Form_Main
    {
        /// <summary>
        /// Required designer variable.
        /// </summary>
        private System.ComponentModel.IContainer components = null;
        protected override void Dispose(bool disposing)
        {
            if (disposing && (components != null))
            {
                components.Dispose();
            }
            base.Dispose(disposing);
        }

        #region Windows Form Designer generated code

        /// <summary>
        /// Required method for Designer support - do not modify
        /// the contents of this method with the code editor.
        /// </summary>
        private void InitializeComponent()
        {
            this.components = new System.ComponentModel.Container();
            System.Windows.Forms.DataVisualization.Charting.ChartArea chartArea1
= new System.Windows.Forms.DataVisualization.Charting.ChartArea();
            System.Windows.Forms.DataVisualization.Charting.Legend legend1 = new
System.Windows.Forms.DataVisualization.Charting.Legend();
            System.Windows.Forms.DataVisualization.Charting.Series series1 = new
System.Windows.Forms.DataVisualization.Charting.Series();
            this.comboBox_Puertos = new System.Windows.Forms.ComboBox();
            this.groupBox_PuertoSerial = new System.Windows.Forms.GroupBox();
            this.button_Refresh = new System.Windows.Forms.Button();
            this.groupBox_Datos = new System.Windows.Forms.GroupBox();
            this.chart_Datos = new
System.Windows.Forms.DataVisualization.Charting.Chart();
```

```
            this.groupBox_Control = new System.Windows.Forms.GroupBox();
            this.button_Stop = new System.Windows.Forms.Button();
            this.button_Iniciar = new System.Windows.Forms.Button();
            this.timer = new System.Windows.Forms.Timer(this.components);
            this.textBox_caracter = new System.Windows.Forms.TextBox();
            this.groupBox1 = new System.Windows.Forms.GroupBox();
            this.label1 = new System.Windows.Forms.Label();
            this.groupBox_PuertoSerial.SuspendLayout();
            this.groupBox_Datos.SuspendLayout();

((System.ComponentModel.ISupportInitialize)(this.chart_Datos)).BeginInit();
            this.groupBox_Control.SuspendLayout();
            this.groupBox1.SuspendLayout();
            this.SuspendLayout();
            //
            // comboBox_Puertos
            //
            this.comboBox_Puertos.FormattingEnabled = true;
            this.comboBox_Puertos.Location = new System.Drawing.Point(6, 19);
            this.comboBox_Puertos.Name = "comboBox_Puertos";
            this.comboBox_Puertos.Size = new System.Drawing.Size(121, 21);
            this.comboBox_Puertos.TabIndex = 0;
            //
            // groupBox_PuertoSerial
            //
            this.groupBox_PuertoSerial.Controls.Add(this.button_Refresh);
            this.groupBox_PuertoSerial.Controls.Add(this.comboBox_Puertos);
            this.groupBox_PuertoSerial.Location = new System.Drawing.Point(12, 12);
            this.groupBox_PuertoSerial.Name = "groupBox_PuertoSerial";
            this.groupBox_PuertoSerial.Size = new System.Drawing.Size(222, 53);
            this.groupBox_PuertoSerial.TabIndex = 1;
            this.groupBox_PuertoSerial.TabStop = false;
            this.groupBox_PuertoSerial.Text = "Puerto Serial Arduino";
            this.groupBox_PuertoSerial.Enter += new
System.EventHandler(this.groupBox_PuertoSerial_Enter);
            //
            // button_Refresh
            //
            this.button_Refresh.Location = new System.Drawing.Point(133, 17);
            this.button_Refresh.Name = "button_Refresh";
            this.button_Refresh.Size = new System.Drawing.Size(75, 23);
            this.button_Refresh.TabIndex = 1;
            this.button_Refresh.Text = "Refresh";
            this.button_Refresh.UseVisualStyleBackColor = true;
            this.button_Refresh.Click += new
System.EventHandler(this.button_Refresh_Click);
            //
```

```
            // groupBox_Datos
            //
            this.groupBox_Datos.Controls.Add(this.chart_Datos);
            this.groupBox_Datos.Location = new System.Drawing.Point(12, 72);
            this.groupBox_Datos.Name = "groupBox_Datos";
            this.groupBox_Datos.Size = new System.Drawing.Size(711, 353);
            this.groupBox_Datos.TabIndex = 2;
            this.groupBox_Datos.TabStop = false;
            this.groupBox_Datos.Text = "Grafica";
            //
            // chart_Datos
            //
            chartArea1.BackColor =
System.Drawing.Color.FromArgb(((int)(((byte)(224)))), ((int)(((byte)(224)))),
((int)(((byte)(224)))));
            chartArea1.Name = "ChartArea1";
            this.chart_Datos.ChartAreas.Add(chartArea1);
            legend1.Enabled = false;
            legend1.Name = "Legend1";
            this.chart_Datos.Legends.Add(legend1);
            this.chart_Datos.Location = new System.Drawing.Point(6, 19);
            this.chart_Datos.Name = "chart_Datos";
            series1.ChartArea = "ChartArea1";
            series1.ChartType =
System.Windows.Forms.DataVisualization.Charting.SeriesChartType.Spline;
            series1.Color = System.Drawing.Color.FromArgb(((int)(((byte)(0)))),
((int)(((byte)(0)))), ((int)(((byte)(192)))));
            series1.IsVisibleInLegend = false;
            series1.Legend = "Legend1";
            series1.Name = "Series1";
            this.chart_Datos.Series.Add(series1);
            this.chart_Datos.Size = new System.Drawing.Size(699, 328);
            this.chart_Datos.TabIndex = 0;
            this.chart_Datos.Text = "Datos del Arduino";
            this.chart_Datos.Click += new
System.EventHandler(this.chart_Datos_Click);
            //
            // groupBox_Control
            //
            this.groupBox_Control.Controls.Add(this.button_Stop);
            this.groupBox_Control.Controls.Add(this.button_Iniciar);
            this.groupBox_Control.Location = new System.Drawing.Point(241, 13);
            this.groupBox_Control.Name = "groupBox_Control";
            this.groupBox_Control.Size = new System.Drawing.Size(190, 53);
            this.groupBox_Control.TabIndex = 3;
            this.groupBox_Control.TabStop = false;
            this.groupBox_Control.Text = "Control de la ejecucion";
```

```
            //
            // button_Stop
            //
            this.button_Stop.Location = new System.Drawing.Point(88, 15);
            this.button_Stop.Name = "button_Stop";
            this.button_Stop.Size = new System.Drawing.Size(75, 23);
            this.button_Stop.TabIndex = 1;
            this.button_Stop.Text = "Stop";
            this.button_Stop.UseVisualStyleBackColor = true;
            this.button_Stop.Click += new
System.EventHandler(this.button_Stop_Click);
            //
            // button_Iniciar
            //
            this.button_Iniciar.Location = new System.Drawing.Point(7, 15);
            this.button_Iniciar.Name = "button_Iniciar";
            this.button_Iniciar.Size = new System.Drawing.Size(75, 23);
            this.button_Iniciar.TabIndex = 0;
            this.button_Iniciar.Text = "Iniciar";
            this.button_Iniciar.UseVisualStyleBackColor = true;
            this.button_Iniciar.Click += new
System.EventHandler(this.button_Iniciar_Click);
            //
            // timer
            //
            this.timer.Tick += new System.EventHandler(this.timer1_Tick);
            //
            // textBox_caracter
            //
            this.textBox_caracter.Location = new System.Drawing.Point(124, 19);
            this.textBox_caracter.Name = "textBox_caracter";
            this.textBox_caracter.Size = new System.Drawing.Size(42, 20);
            this.textBox_caracter.TabIndex = 1;
            //
            // groupBox1
            //
            this.groupBox1.Controls.Add(this.textBox_caracter);
            this.groupBox1.Location = new System.Drawing.Point(438, 13);
            this.groupBox1.Name = "groupBox1";
            this.groupBox1.Size = new System.Drawing.Size(285, 52);
            this.groupBox1.TabIndex = 4;
            this.groupBox1.TabStop = false;
            this.groupBox1.Text = "Caracter Obtenido";
            //
            // label1
            //
            this.label1.AutoSize = true;
```

```
            this.label1.Location = new System.Drawing.Point(12, 442);
            this.label1.Name = "label1";
            this.label1.Size = new System.Drawing.Size(139, 13);
            this.label1.TabIndex = 5;
            this.label1.Text = "Christie Nucamendi Canelas";
            // 
            // Form_Main
            // 
            this.AutoScaleDimensions = new System.Drawing.SizeF(6F, 13F);
            this.AutoScaleMode = System.Windows.Forms.AutoScaleMode.Font;
            this.BackColor = System.Drawing.Color.FromArgb(((int)(((byte)(192)))), 
((int)(((byte)(192)))), ((int)(((byte)(255)))));
            this.ClientSize = new System.Drawing.Size(735, 497);
            this.Controls.Add(this.label1);
            this.Controls.Add(this.groupBox1);
            this.Controls.Add(this.groupBox_Control);
            this.Controls.Add(this.groupBox_Datos);
            this.Controls.Add(this.groupBox_PuertoSerial);
            this.MaximizeBox = false;
            this.Name = "Form_Main";
            this.Text = "Prototipo de residencia";
            this.FormClosed += new 
System.Windows.Forms.FormClosedEventHandler(this.Form_Main_FormClosed);
            this.Load += new System.EventHandler(this.Form_Main_Load);
            this.groupBox_PuertoSerial.ResumeLayout(false);
            this.groupBox_Datos.ResumeLayout(false);
            ((System.ComponentModel.ISupportInitialize)(this.chart_Datos)).EndInit();
            this.groupBox_Control.ResumeLayout(false);
            this.groupBox1.ResumeLayout(false);
            this.groupBox1.PerformLayout();
            this.ResumeLayout(false);
            this.PerformLayout();

        }

        #endregion

        private System.Windows.Forms.ComboBox comboBox_Puertos;
        private System.Windows.Forms.GroupBox groupBox_PuertoSerial;
        private System.Windows.Forms.Button button_Refresh;
        private System.Windows.Forms.GroupBox groupBox_Datos;
        private System.Windows.Forms.GroupBox groupBox_Control;
        private System.Windows.Forms.Button button_Iniciar;
        private System.Windows.Forms.Button button_Stop;
        private System.Windows.Forms.DataVisualization.Charting.Chart 
chart_Datos;
        private System.Windows.Forms.Timer timer;
```

```
        private System.Windows.Forms.TextBox textBox_caracter;
        private System.Windows.Forms.GroupBox groupBox1;
        private System.Windows.Forms.Label label1;
    }
}
```

4.2 Resultados

En esta parte se encontraran los resultados de la pruebas que se hicieron para cada sensor, con la finalidad de ver su comportamiento.

4.2.1 Sensor de corriente

.

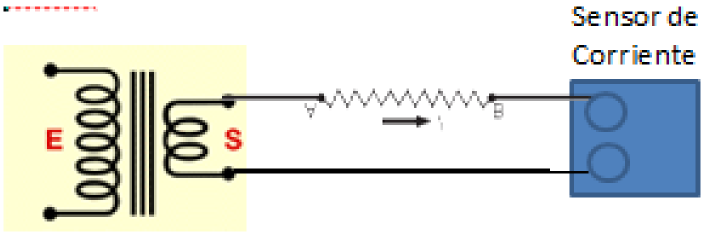


Figura 4.1 Caracterización del Sensor de Corriente

En este diagrama podemos observa cómo fue conectado el sensor de corriente, se usa un transformador para poder reducir la señal de entrada a un voltaje de 12v y se hacen mediciones de acuerdo con una variación de Resistencia para poder ver el comportamiento encada una de la corriente que está en ese instante y el voltaje que encontramos de salida con respecto a la corriente que tenemos. Para corroborarlo se muestra la siguiente tabla con los resultados obtenidos.

Mediciones	Voltaje de salida del Transformador (Vrms)	Resistencia OHMS	Corriente AMPERES (Irms)	Voltaje de salida del sensor (mV)
1	12	12	1	260
2	12	7	1.7143	370
3	12	5.6	2.1429	520
4	12	5	2.4	580

Tabla 4.1 Voltajes de salida del Sensor de Corriente

También se adjunta una grafica para especificar la variación del voltaje del sensor con respecto a la corriente.

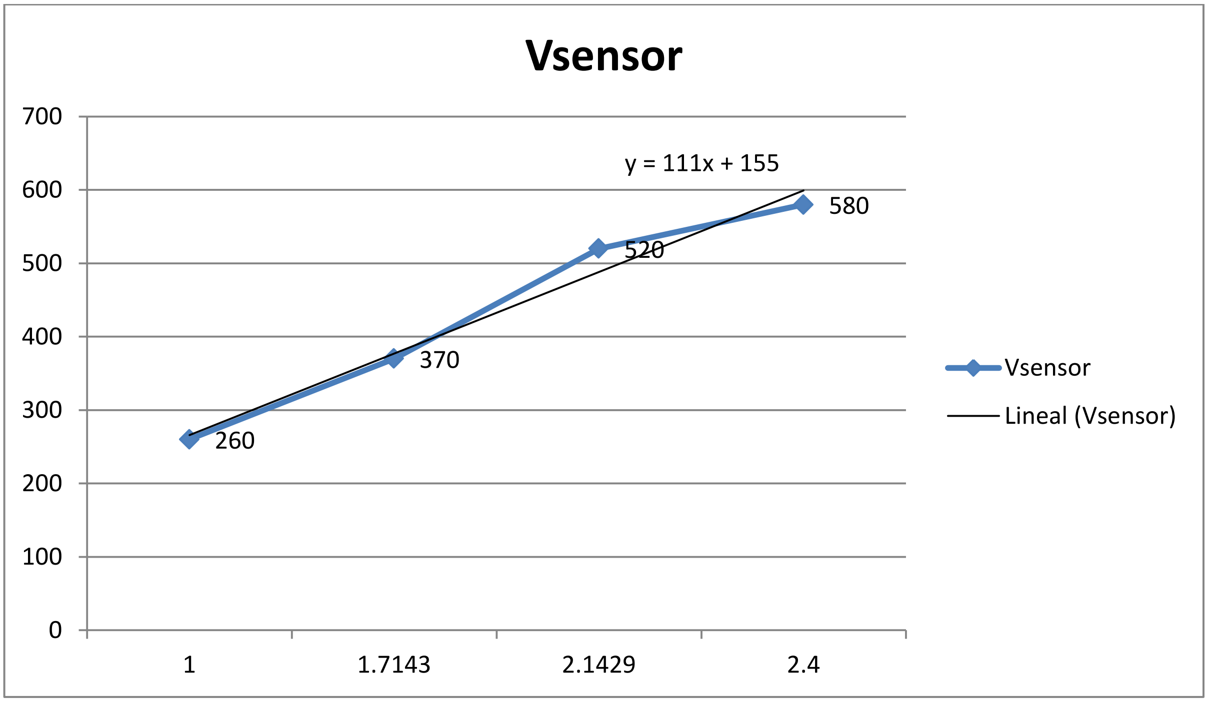
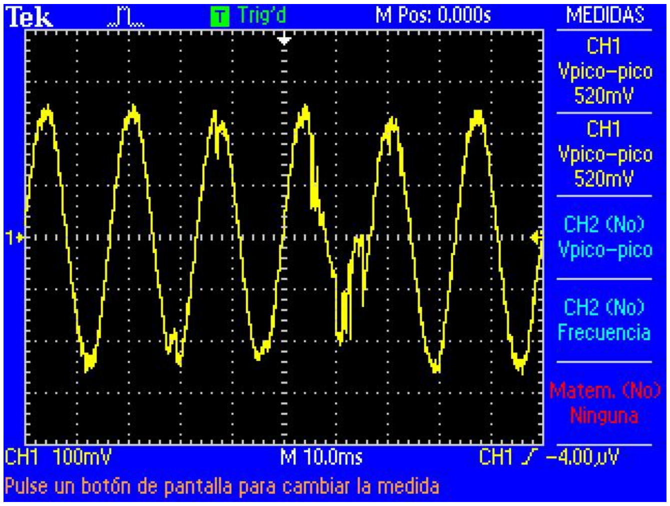


Figura 4.2 Grafica de Valores y ecuación que describe al Sensor

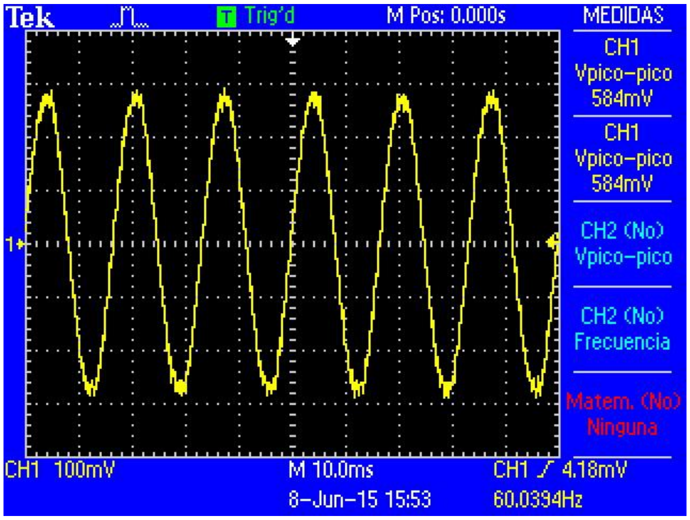
CAPTURAS DE OSCILOGRAMAS DEL CIRCUITO.

Se muestran las capturas del circuito con el voltaje de salida del sensor para cada cambio de resistencia que se efectuó.

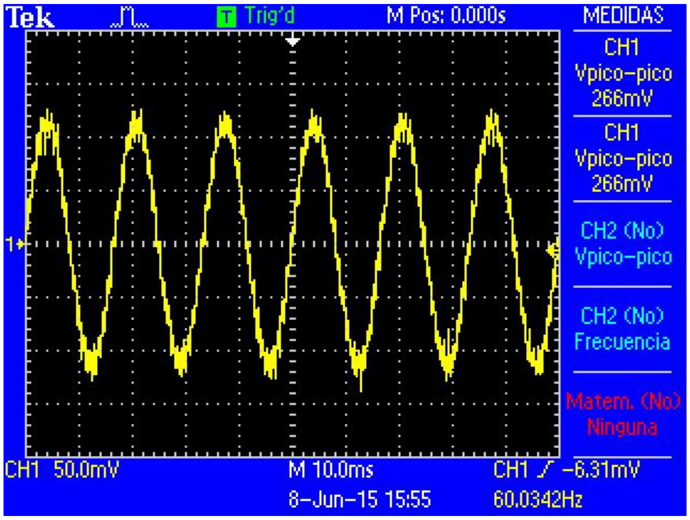


Pulse un botón de pantalla para cambiar la medida

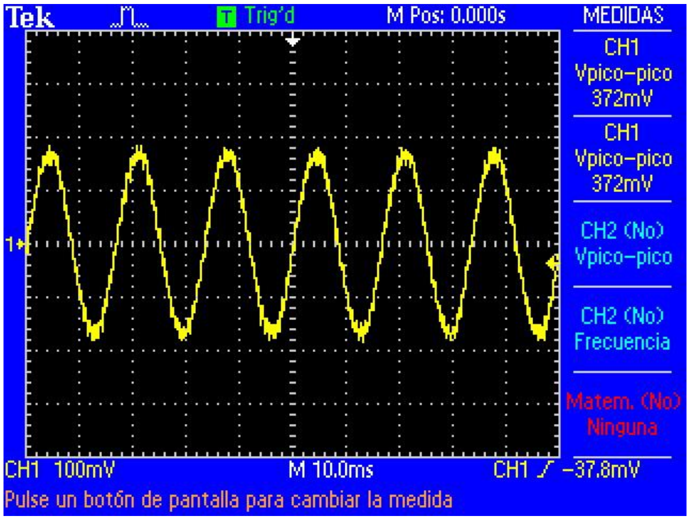
Captura 1. Donde se muestra el voltaje de salida de 520mV cuando se tiene una resistencia de 5.6Ω, con un voltaje de entrada del transformador de 12v, obteniendo una corriente de 2.1429A.



Captura2. Se muestra el voltaje de salida de 584mV cuando se tiene una resistencia de 5Ω, con un voltaje de entrada del transformador de 12v, obteniendo una corriente de 2.4A.



Captura 3. Se muestra el voltaje de salida de 266mV cuando se tiene una resistencia de 12Ω, con un voltaje de entrada del transformador de 12v, obteniendo una corriente de 1A.



Captura 4. Se muestra el voltaje de salida de 372mV cuando se tiene una resistencia de 7Ω, con un voltaje de entrada del transformador de 12v, obteniendo una corriente de 1.7143A.

4.2.2 Resultado del circuito integrado LCT1963 con acoplamiento RMS – DC

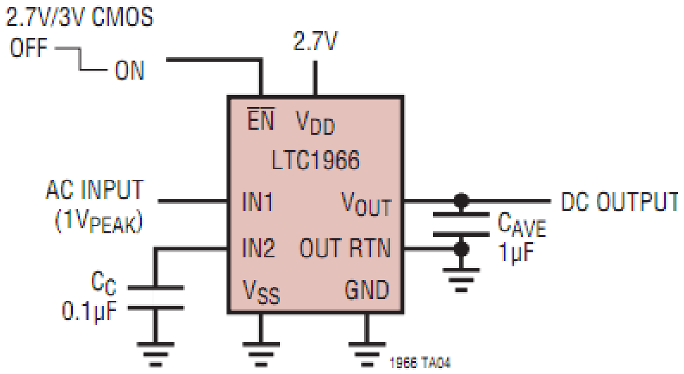


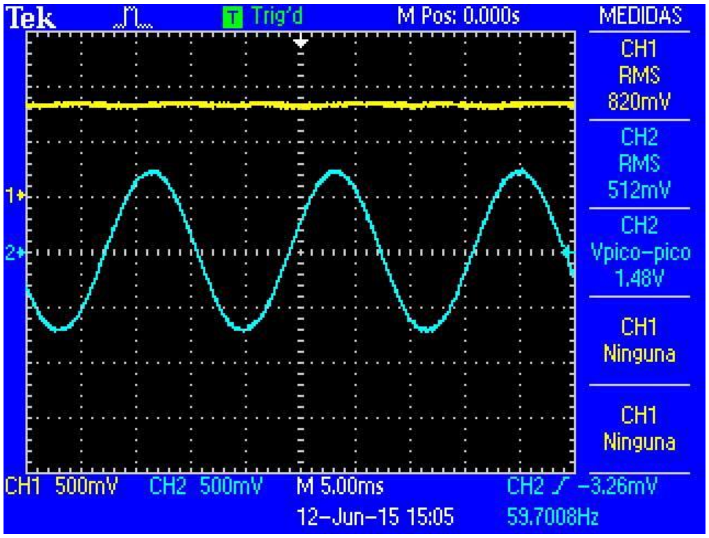
Figura 4.3 Diagrama utilizado.

Con este sensor se hicieron los cálculos para obtener el voltaje Vrms de entrada y el voltaje pico a pico de entrada con respecto al Vrms del circuito integrado.

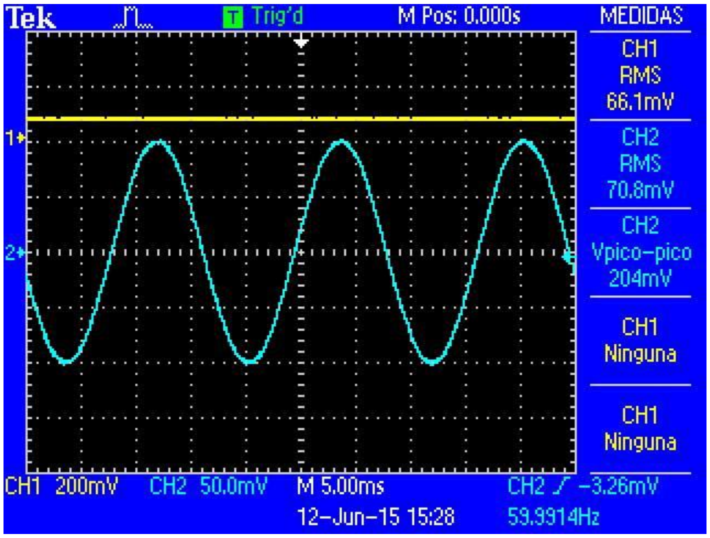
Voltaje pico-pico (entrada)	Vrms (entrada)	Vrms de circuito integrado
1.48V	513 mV	818mV
1.42V	490 mV	757 mV
1.34V	459 mV	635 mV
1.24V	427 mV	591 mV
1.10V	375 mV	530 mV
1.02V	351 mV	450 mV
920mV	312 mV	318 mV
820 mV	277 mV	287 mV
720 mV	240 mV	252 mV
608 mV	211 mV	200 mV
512 mV	179 mV	172 mV
400 mV	139 mV	134 mV
304 mV	106 mV	102 mV
200 mV	70.9 mV	66.1 mV
104 mV	35.8 mV	37.7 mV

Tabla 4.2 Resultados de las mediciones

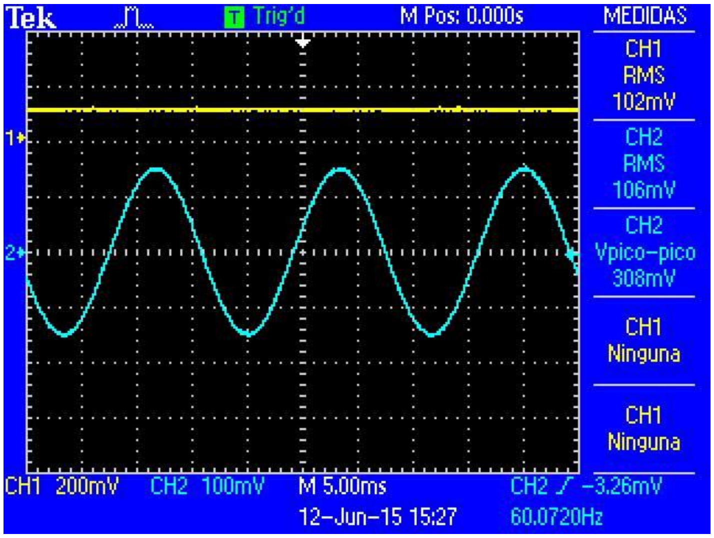
CAPTURA DEL CIRCUITO



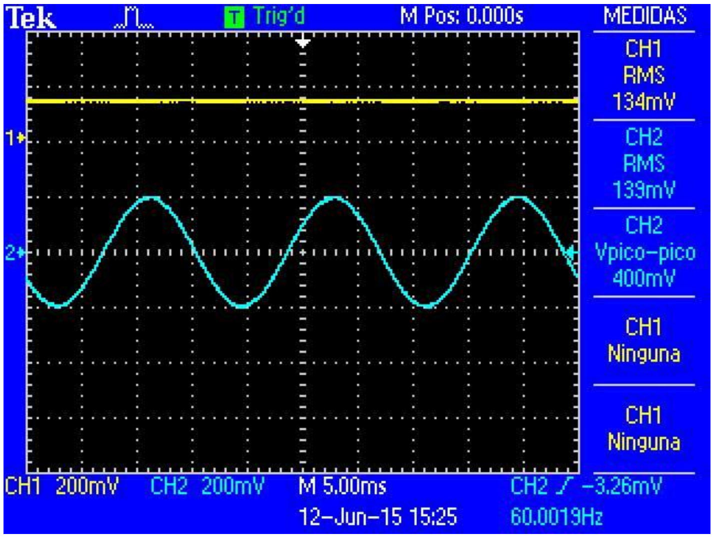
Captura 1. Esta imagen es del Vrms del circuito integrado de 818mV con un Vrms de entrada de 513 mV a un voltaje pico a pico de 1.48v.



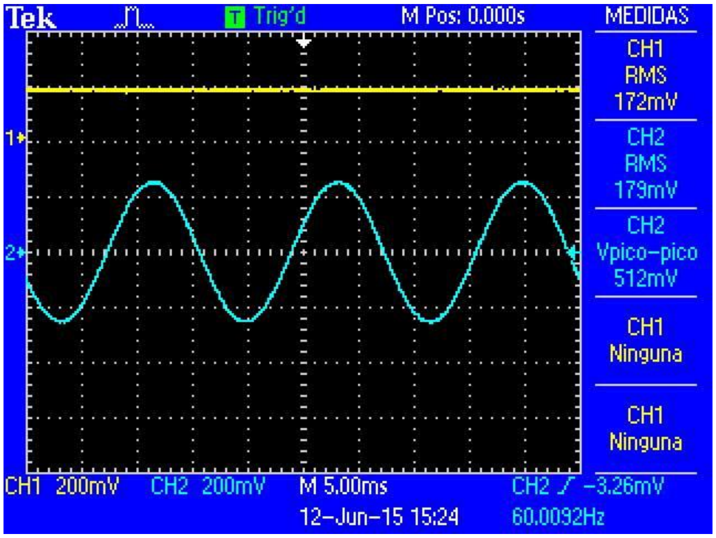
Captura 2. Esta imagen es del Vrms del circuito integrado de 66.1mV con un Vrms de entrada de 70.9mV a un voltaje pico a pico de 200mV.



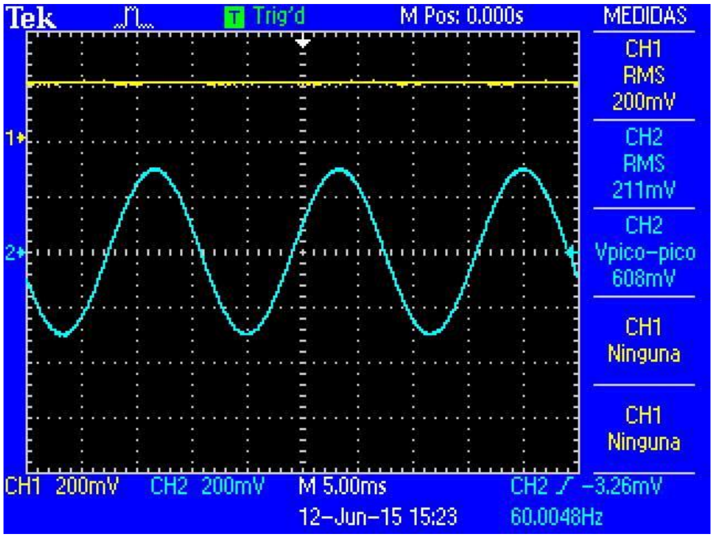
Captura 3. Esta imagen es del Vrms del circuito integrado de 102mV con un Vrms de entrada de 106mV a un voltaje pico a pico de 304mV.



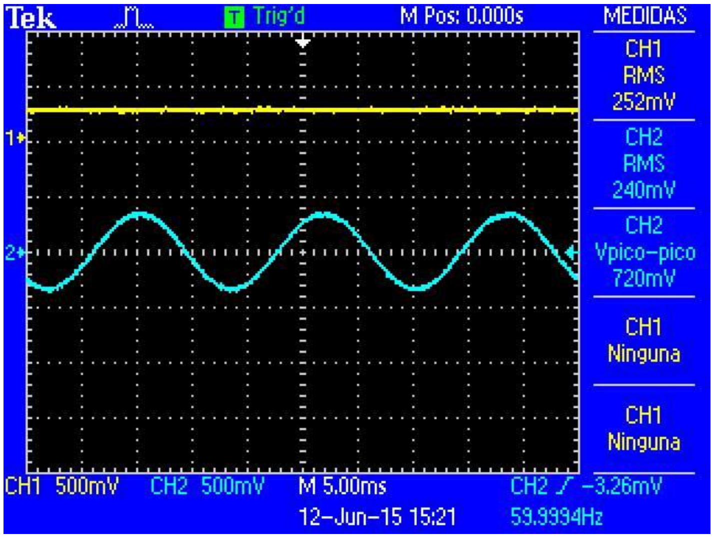
Captura 4. Esta imagen es del Vrms del circuito integrado de 134mV con un Vrms de entrada de 139mV a un voltaje pico a pico de 400mV.



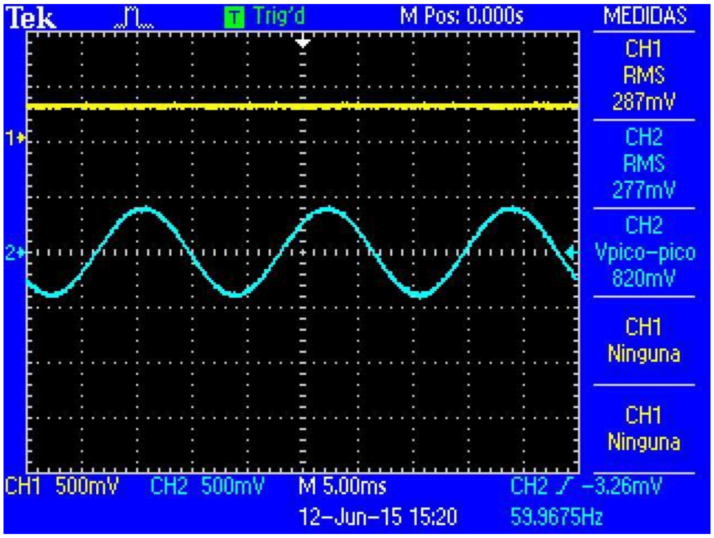
Captura 5. Esta imagen es del Vrms del circuito integrado de 172mV con un Vrms de entrada de 179mV a un voltaje pico a pico de 512mV.



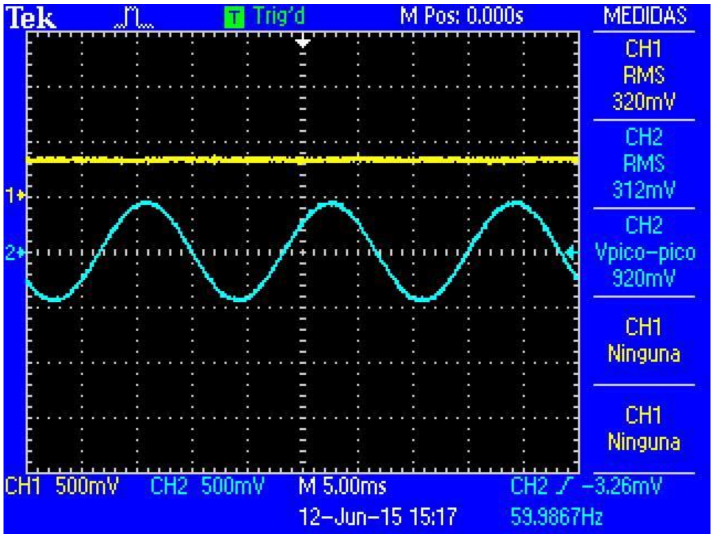
Captura 6. Esta imagen es del Vrms del circuito integrado de 200mV con un Vrms de entrada de 211mV a un voltaje pico a pico de 608mV.



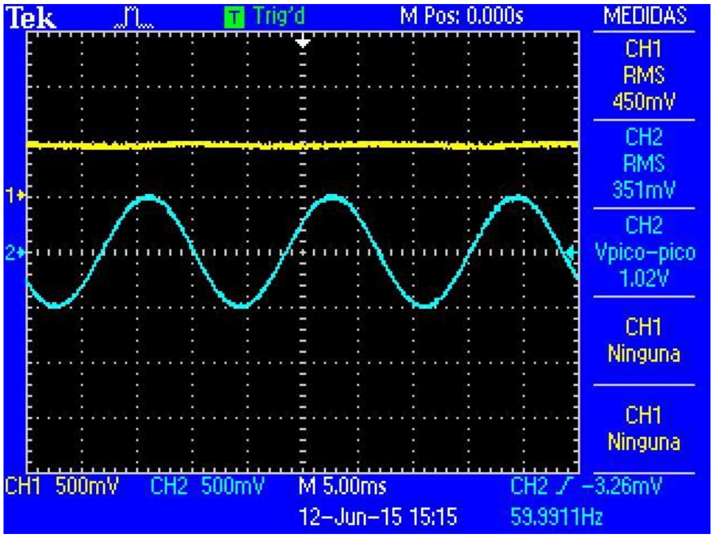
Captura 7. Esta imagen es del Vrms del circuito integrado de 252mV con un Vrms de entrada de 240mV a un voltaje pico a pico de 720mV.



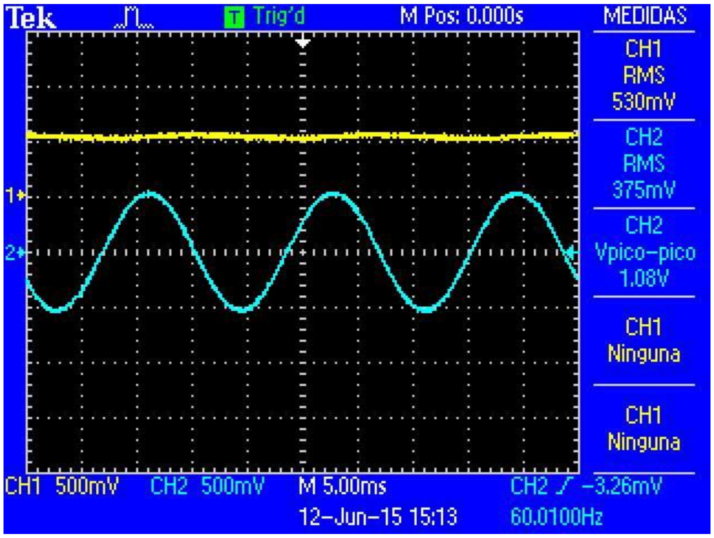
Captura 8. Esta imagen es del Vrms del circuito integrado de 287mV con un Vrms de entrada de 277mV a un voltaje pico a pico de 820mV.



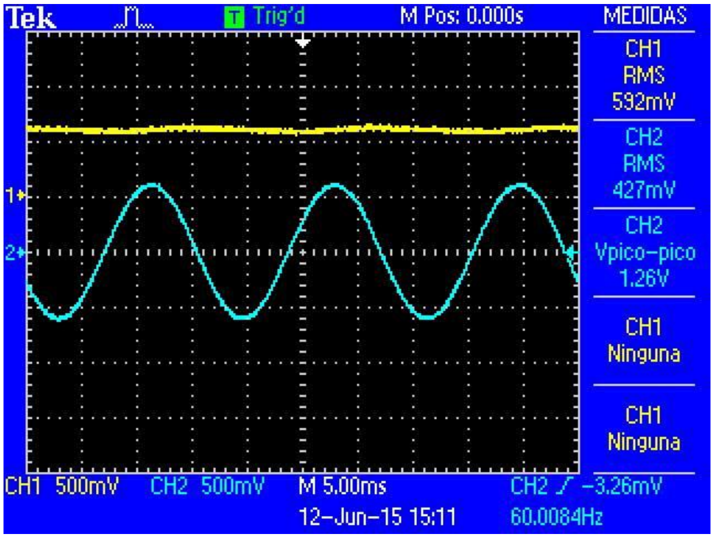
Captura 9. Esta imagen es del Vrms del circuito integrado de 318mV con un Vrms de entrada de 312mV a un voltaje pico a pico de 920mV.



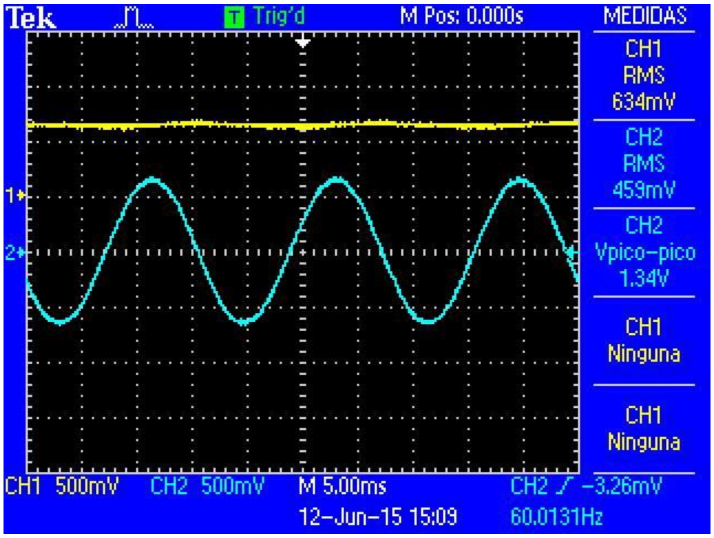
Captura 10. Esta imagen es del Vrms del circuito integrado de 450mV con un Vrms de entrada de 351mV a un voltaje pico a pico de 1.02V.



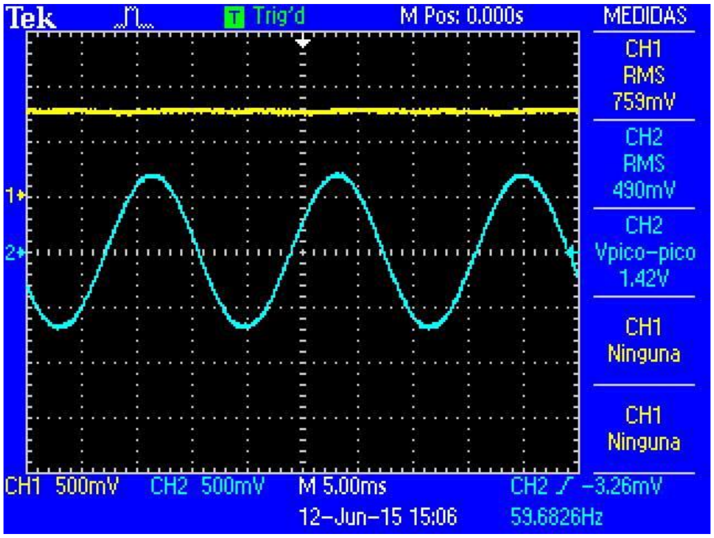
Captura 11. Esta imagen es del Vrms del circuito integrado de 530mV con un Vrms de entrada de 357mV a un voltaje pico a pico de 1.10V.



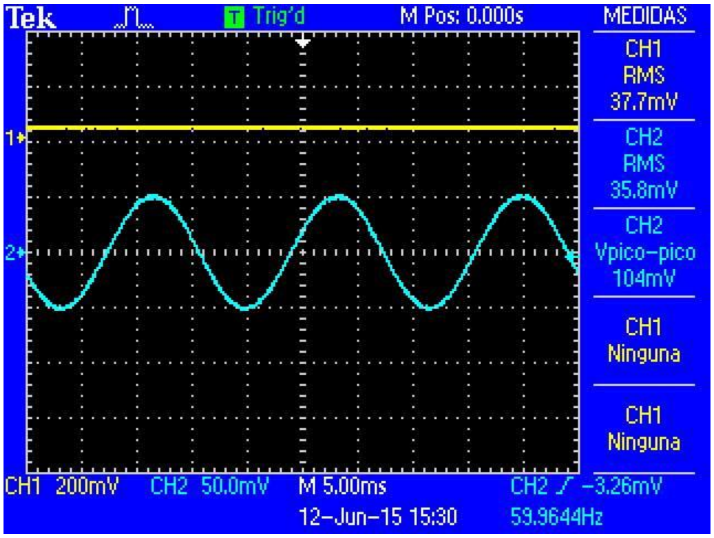
Captura 12. Esta imagen es del Vrms del circuito integrado de 591mV con un Vrms de entrada de 427mV a un voltaje pico a pico de 1.24V.



Captura 13. Esta imagen es del Vrms del circuito integrado de 635mV con un Vrms de entrada de 459mV a un voltaje pico a pico de 1.34V.



Captura 14. Esta imagen es del Vrms del circuito integrado de 757mV con un Vrms de entrada de 490mV a un voltaje pico a pico de 1.42V.



Captura 15. Esta imagen es del Vrms del circuito integrado de 37.7mV con un Vrms de entrada de 35.8mV a un voltaje pico a pico de 104mV.

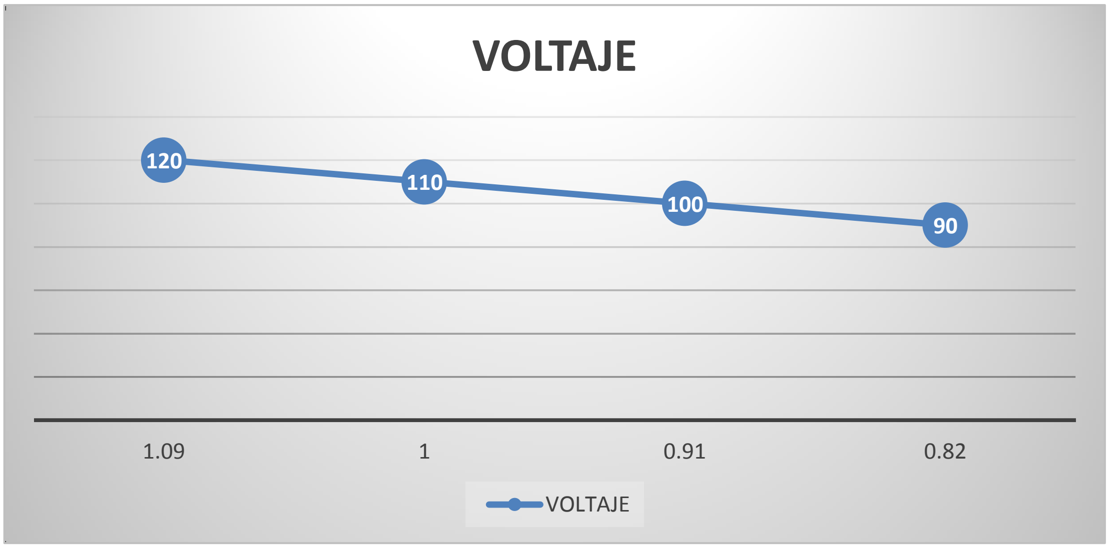
4.2.3 Resultado del voltaje de salida del sensor del voltaje pico.

En esta tabla se pueden observar los resultados de voltaje de entradas de corriente alterna, la cual se fue variando para poder conocer si existía una variación de voltaje que de salida y los siguientes cálculos fueron los resultados.

VOLTAJE DE ENTRADA	VOLTAJE DE SALIDA
120VAC	1.09VAC
110VAC	1.00VAC
100VAC	0.91VAC
90VAC	0.82VAC

Tabla 4.3 Voltajes de salida del sensor de voltaje pico

Se adjunta también una grafica para poder observar como al disminuir el voltaje de entrada de CA también podeos observar una disminución del voltaje de salida como se muestra en la siguiente grafica

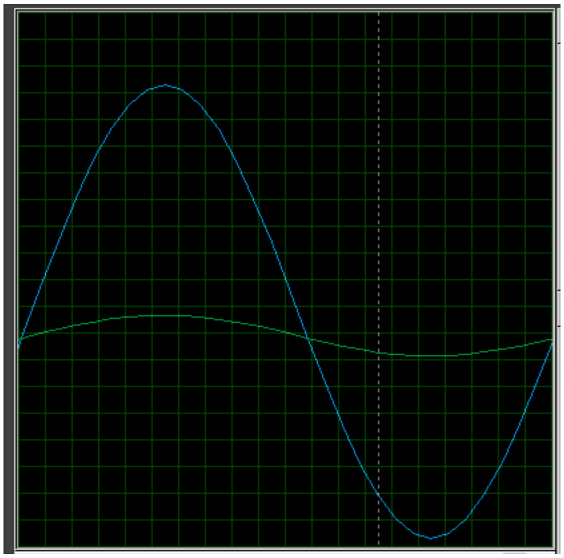


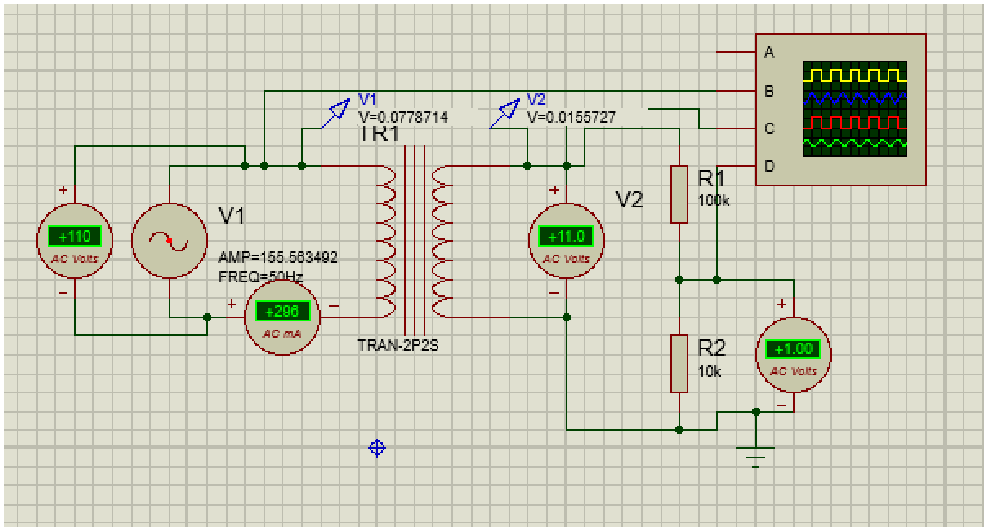
4.4 graficas obtenidas con los datos de la tabla.

CAPTURA DEL CIRCUITO.

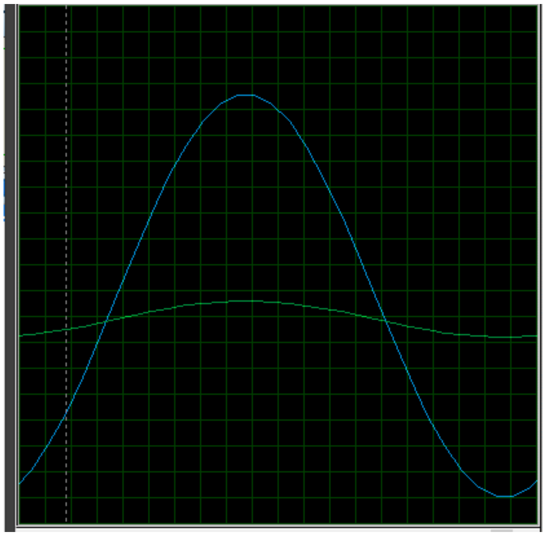
Para las siguientes imágenes vamos a observar el comportamiento de la onda para el voltaje de entrada y el voltaje de salida y se adjunta la imagen de los datos en el osciloscopio.

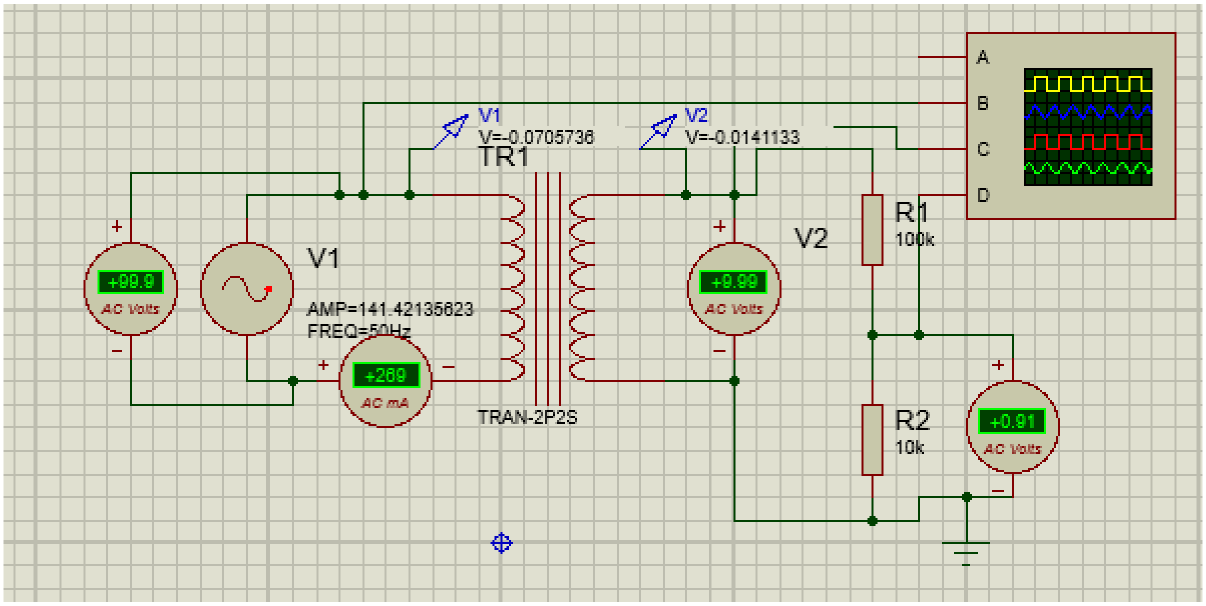
Captura 1. Entrada de 110 Vca con salida de 1Vca.



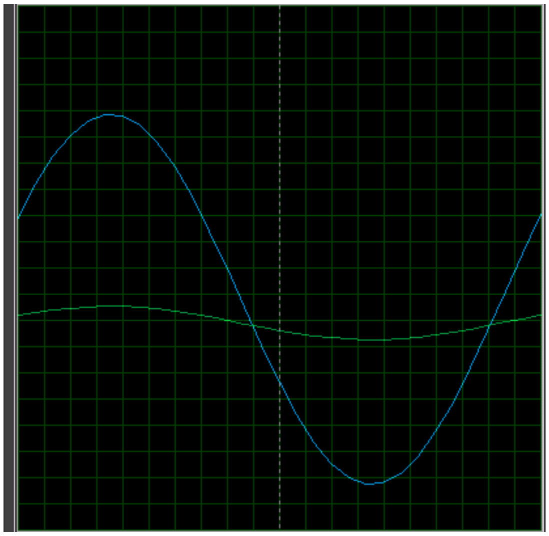


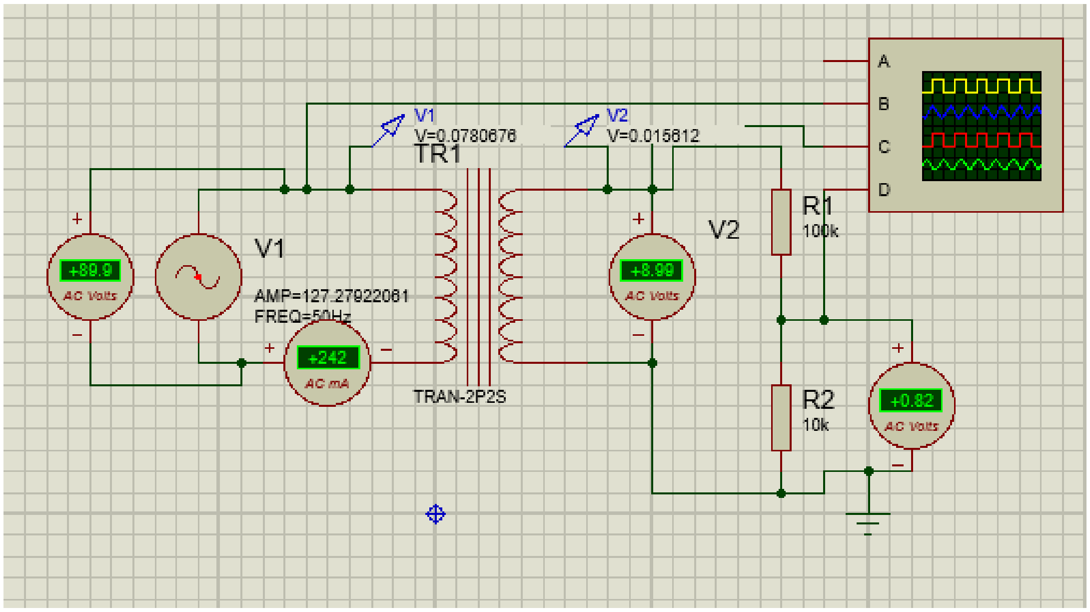
Captura 2. Entrada de 100 Vca con salida de 0.91Vca



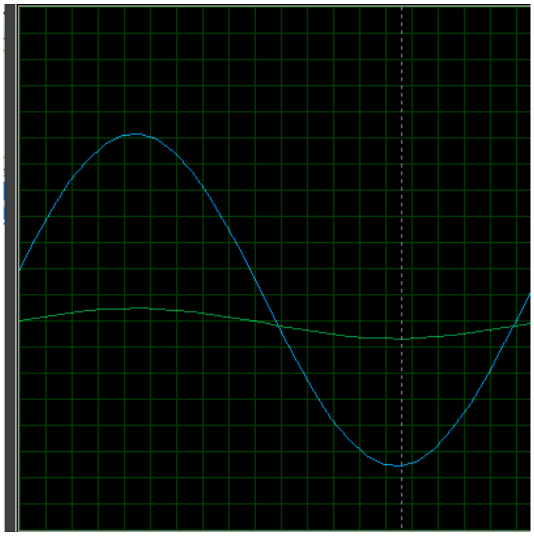


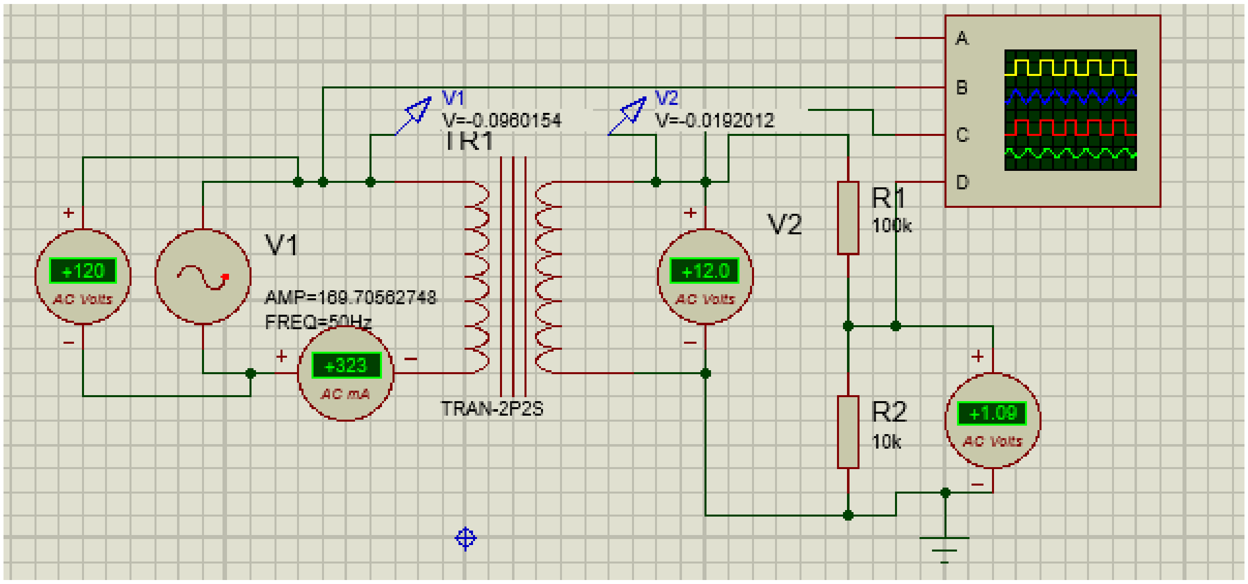
Captura 3. Entrada de 90 Vca con salida de 0.82Vca





Captura 4. Entrada de 120 Vca con salida de 1.09Vca




A
B
C
D
V1
V=-0.0960154
V2
V=-0.0192012
TR1
+
+120
AC Volts
−
V1
AMP=169.70562748
FREQ=50Hz
+
+323
AC mA
−
TRAN-2P2S
+
+12.0
AC Volts
−
V2
R1
100k
R2
10k
+
+1.09
AC Volts
−

CONCLUSIONES.

De manera idónea se han alcanzado la mayoría de los objetivos planteados durante el desarrollo del proyecto tomando como punto de referencia las necesidades que existen en la industria y la implementación de los DVR, se comenzó a diseñar un programa basado en entorno visual de alto nivel para poder monitorear gráficamente el estado de los sensores de la etapa de protección del mismo DVR, así mismo, se llevó a cabo el diseño e implementación de los prototipos de instrumentación de monitoreo y censado para la protección del mismo DVR, utilizando sensores muy específicos para medir Corriente y voltaje Rms , con la finalidad de poder apreciar las distorsiones propias del sistema en una línea AC, que por sus propiedades naturales posee variaciones o ruidos que pueden afectar un sistema, continuo a esto se llevó a cabo una etapa de transformación de la señal AC a DC con la finalidad de poder ingresar al controlador, en este caso en particular el controlador en la plataforma Arduino y así realizar en enlace para su monitoreo en tiempo real.

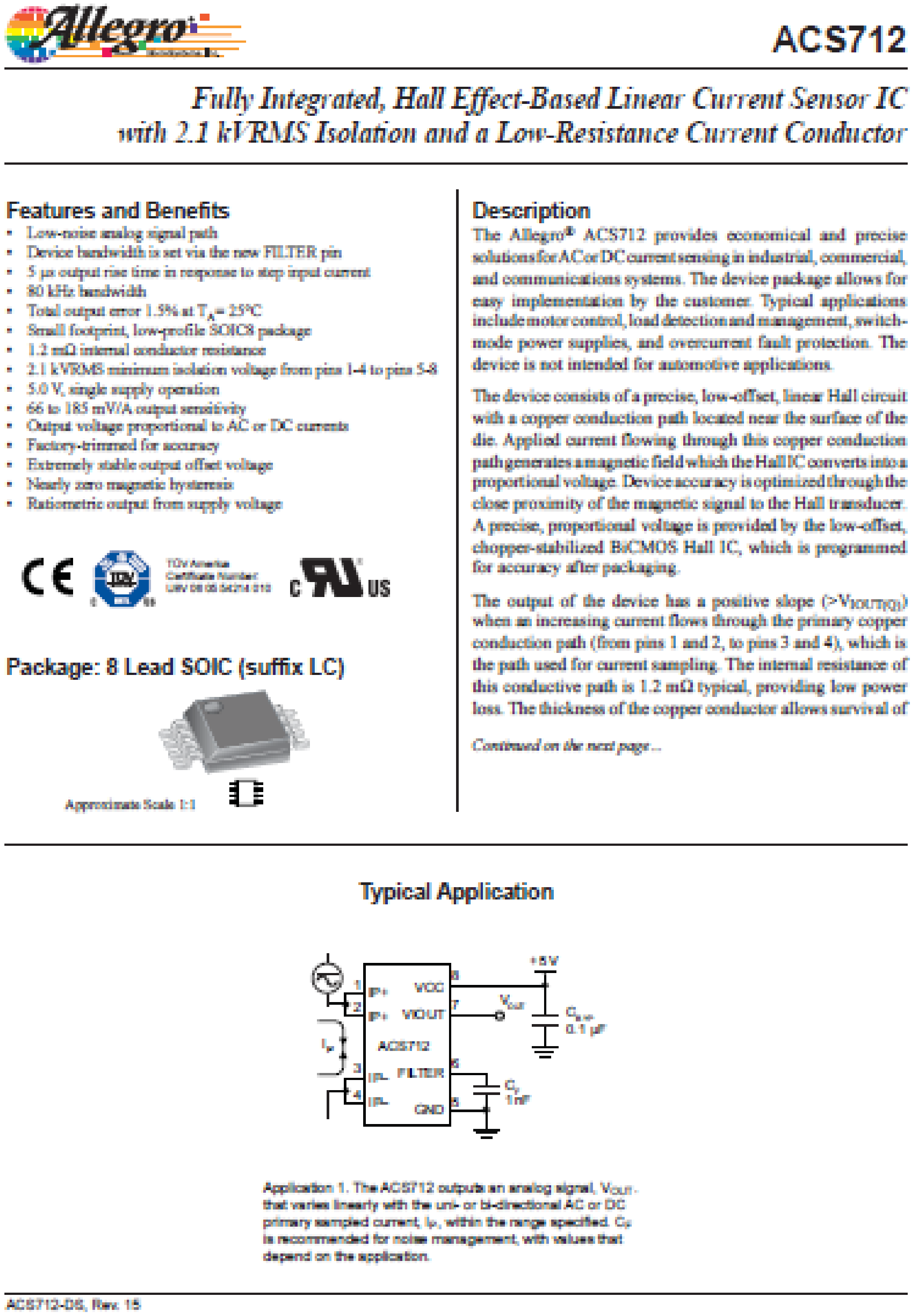
Dado al desarrollo del proyecto y a los resultados que se mostraron en los anexos y en el apartado de resultados, se puede decir que se alcanzaron la mayoría de los objetivos propuestos, sin embargo, hace falta aún mucho desarrollo en esta materia ya que es un tema de complejidad y elongación alta, sin embargo la etapa de protección ya es un prototipo que se puede y se pretende que se siga perfeccionando.

REFERENCIA.

Se realizaron las investigaciones del trabajo con ayuda de páginas de internet como fue de mucha ayuda Google. El libro de amplificadores operacionales (Driscoll, 1999) nos fue de mucha ayuda para ver el diagrama de uno de los componentes fundamentales para el circuito. La pagina 5Hz nos proporciono los datos específicos de los sensores de corriente (Electronica, 2015). Otras páginas que nos sirvieron para las especificaciones del Arduino como (ohms, 2015). Datashet se sensores en (Allegro MicroSystems) y (Olea, 2001) que fueron útiles para el complemento de nuestra investigación.

ANEXOS

Anexo A. Sensor de Corriente



ACS712

Fully Integrated, Hall Effect-Based Linear Current Sensor IC with 2.1 kVRMS Isolation and a Low-Resistance Current Conductor

Features and Benefits

- Low-noise analog signal path
- Device bandwidth is set via the new FILTER pin
- 5 μs output rise time in response to step input current
- 80 kHz bandwidth
- Total output error 1.5% at T_A = 25°C
- Small footprint, low-profile SOIC8 package
- 1.2 mΩ internal conductor resistance
- 2.1 kVRMS minimum isolation voltage from pins 1-4 to pins 5-8
- 5.0 V, single supply operation
- 66 to 185 mV/A output sensitivity
- Output voltage proportional to AC or DC currents
- Factory-trimmed for accuracy
- Extremely stable output offset voltage
- Nearly zero magnetic hysteresis
- Ratiometric output from supply voltage

TÜV America
Certificate Number:
U8V 06 05 54214 010

Package: 8 Lead SOIC (suffix LC)

Approximate Scale 1:1

Description

The Allegro® ACS712 provides economical and precise solutions for AC or DC current sensing in industrial, commercial, and communications systems. The device package allows for easy implementation by the customer. Typical applications include motor control, load detection and management, switch-mode power supplies, and overcurrent fault protection. The device is not intended for automotive applications.

The device consists of a precise, low-offset, linear Hall circuit with a copper conduction path located near the surface of the die. Applied current flowing through this copper conduction path generates a magnetic field which the Hall IC converts into a proportional voltage. Device accuracy is optimized through the close proximity of the magnetic signal to the Hall transducer. A precise, proportional voltage is provided by the low-offset, chopper-stabilized BiCMOS Hall IC, which is programmed for accuracy after packaging.

The output of the device has a positive slope (>$V_{IOUT(Q)}$) when an increasing current flows through the primary copper conduction path (from pins 1 and 2, to pins 3 and 4), which is the path used for current sampling. The internal resistance of this conductive path is 1.2 mΩ typical, providing low power loss. The thickness of the copper conductor allows survival of

Continued on the next page...

Typical Application

Application 1. The ACS712 outputs an analog signal, V_{OUT}, that varies linearly with the uni- or bi-directional AC or DC primary sampled current, I_P, within the range specified. C_F is recommended for noise management, with values that depend on the application.

ACS712-DS, Rev. 15

Description (continued)

the device at up to 5× overcurrent conditions. The terminals of the conductive path are electrically isolated from the signal leads (pins 5 through 8). This allows the ACS712 to be used in applications requiring electrical isolation without the use of opto-isolators or other costly isolation techniques.

The ACS712 is provided in a small, surface mount SOIC8 package. The leadframe is plated with 100% matte tin, which is compatible with standard lead (Pb) free printed circuit board assembly processes. Internally, the device is Pb-free, except for flip-chip high-temperature Pb-based solder balls, currently exempt from RoHS. The device is fully calibrated prior to shipment from the factory.

Selection Guide

Part Number	Packing*	T_A (°C)	Optimized Range, I_P (A)	Sensitivity, Sens (Typ) (mV/A)
ACS712ELCTR-05B-T	Tape and reel, 3000 pieces/reel	–40 to 85	±5	185
ACS712ELCTR-20A-T	Tape and reel, 3000 pieces/reel	–40 to 85	±20	100
ACS712ELCTR-30A-T	Tape and reel, 3000 pieces/reel	–40 to 85	±30	66

*Contact Allegro for additional packing options.

Absolute Maximum Ratings

Characteristic	Symbol	Notes	Rating	Units
Supply Voltage	V_{CC}		8	V
Reverse Supply Voltage	V_{RCC}		–0.1	V
Output Voltage	V_{IOUT}		8	V
Reverse Output Voltage	V_{RIOUT}		–0.1	V
Output Current Source	$I_{IOUT(Source)}$		3	mA
Output Current Sink	$I_{IOUT(Sink)}$		10	mA
Overcurrent Transient Tolerance	I_P	1 pulse, 100 ms	100	A
Nominal Operating Ambient Temperature	T_A	Range E	–40 to 85	°C
Maximum Junction Temperature	T_J(max)		165	°C
Storage Temperature	T_{stg}		–65 to 170	°C

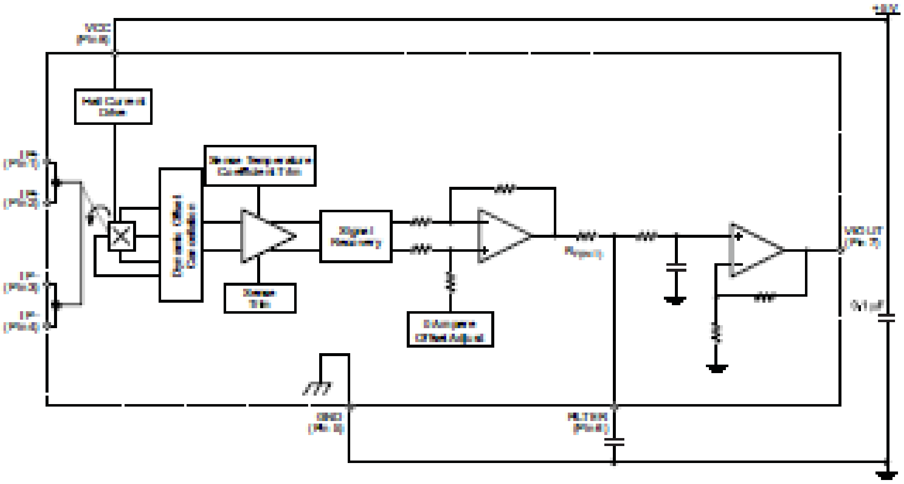
Isolation Characteristics

Characteristic	Symbol	Notes	Rating	Unit
Dielectric Strength Test Voltage*	V_{ISO}	Agency type-tested for 60 seconds per UL standard 60950-1, 1st Edition	2100	VAC
Working Voltage for Basic Isolation	V_{WFSI}	For basic (single) isolation per UL standard 60950-1, 1st Edition	354	VDC or V_{pk}
Working Voltage for Reinforced Isolation	V_{WFRI}	For reinforced (double) isolation per UL standard 60950-1, 1st Edition	184	VDC or V_{pk}

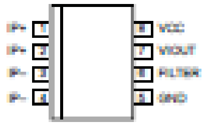
* Allegro does not conduct 60-second testing. It is done only during the UL certification process.

Parameter	Specification
Fire and Electric Shock	CAN/CSA-C22.2 No. 60950-1-03 UL 60950-1:2003 EN 60950-1:2001

Functional Block Diagram



Pin-out Diagram



Terminal List Table

Number	Name	Description
1 and 2	IP+	Terminals for current being sampled; fused internally
3 and 4	IP−	Terminals for current being sampled; fused internally
5	GND	Signal ground terminal
6	FILTER	Terminal for external capacitor that sets bandwidth
7	VIOUT	Analog output signal
8	VCC	Device power supply terminal

COMMON OPERATING CHARACTERISTICS[1] over full range of T_A, C_F = 1 nF, and V_{CC} = 5 V, unless otherwise specified

Characteristic	Symbol	Test Conditions	Min.	Typ.	Max.	Units
ELECTRICAL CHARACTERISTICS						
Supply Voltage	V_{CC}		4.5	5.0	5.5	V
Supply Current	I_{CC}	V_{CC} = 5.0 V, output open	–	10	13	mA
Output Capacitance Load	C_{LOAD}	VIOUT to GND	–	–	10	nF
Output Resistive Load	R_{LOAD}	VIOUT to GND	4.7	–	–	kΩ
Primary Conductor Resistance	$R_{PRIMARY}$	T_A = 25°C	–	1.2	–	mΩ
Rise Time	t_r	$I_P = I_P(max)$, T_A = 25°C, C_{OUT} = open	–	3.5	–	μs
Frequency Bandwidth	f	–3 dB, T_A = 25°C; I_P is 10 A peak-to-peak	–	80	–	kHz
Nonlinearity	E_{LIN}	Over full range of I_P	–	1.5	–	%
Symmetry	E_{SYM}	Over full range of I_P	98	100	102	%
Zero Current Output Voltage	$V_{IOUT(Q)}$	Bidirectional; I_P = 0 A, T_A = 25°C	–	V_{CC} × 0.5	–	V
Power-On Time	t_{PO}	Output reaches 90% of steady-state level, T_J = 25°C, 20 A present on leadframe	–	35	–	μs
Magnetic Coupling[2]			–	12	–	G/A
Internal Filter Resistance[3]	$R_{F(INT)}$			1.7		kΩ

[1] Device may be operated at higher primary current levels, I_P, and ambient, T_A, and internal leadframe temperatures, T_A, provided that the Maximum Junction Temperature, T_J(max), is not exceeded.

[2] 1G = 0.1 mT.

[3] $R_{F(INT)}$ forms an RC circuit via the FILTER pin.

COMMON THERMAL CHARACTERISTICS[1]

			Min.	Typ.	Max.	Units
Operating Internal Leadframe Temperature	T_A	E range	–40	–	85	°C
					Value	Units
Junction-to-Lead Thermal Resistance[2]	$R_{θJL}$	Mounted on the Allegro ASEK 712 evaluation board			5	°C/W
Junction-to-Ambient Thermal Resistance	$R_{θJA}$	Mounted on the Allegro 85-0322 evaluation board, includes the power consumed by the board			23	°C/W

[1] Additional thermal information is available on the Allegro website.

[2] The Allegro evaluation board has 1500 mm^2 of 2 oz. copper on each side, connected to pins 1 and 2, and to pins 3 and 4, with thermal vias connecting the layers. Performance values include the power consumed by the PCB. Further details on the board are available from the Frequently Asked Questions document on our website. Further information about board design and thermal performance also can be found in the Applications Information section of this datasheet.

x05B PERFORMANCE CHARACTERISTICS[1] T_A = –40°C to 85°C, C_F = 1 nF, and V_{CC} = 5 V, unless otherwise specified

Characteristic	Symbol	Test Conditions	Min.	Typ.	Max.	Units
Optimized Accuracy Range	I_P		–5	–	5	A
Sensitivity	Sens	Over full range of I_P, T_A = 25°C	180	185	190	mV/A
Noise	$V_{NOISE(PP)}$	Peak-to-peak, T_A = 25°C, 185 mV/A programmed Sensitivity, C_F = 47 nF, C_{OUT} = open, 2 kHz bandwidth	–	21	–	mV
Zero Current Output Slope	$\Delta V_{OUT(Q)}$	T_A = –40°C to 25°C	–	–0.26	–	mV/°C
		T_A = 25°C to 150°C	–	–0.08	–	mV/°C
Sensitivity Slope	∆Sens	T_A = –40°C to 25°C	–	0.054	–	mV/A/°C
		T_A = 25°C to 150°C	–	–0.008	–	mV/A/°C
Total Output Error[2]	E_{TOT}	I_P =±5 A, T_A = 25°C	–	±1.5	–	%

[1] Device may be operated at higher primary current levels, I_P, and ambient temperatures, T_A, provided that the Maximum Junction Temperature, $T_{J(max)}$, is not exceeded.

[2] Percentage of I_P, with I_P = 5 A. Output filtered.

x20A PERFORMANCE CHARACTERISTICS[1] T_A = –40°C to 85°C, C_F = 1 nF, and V_{CC} = 5 V, unless otherwise specified

Characteristic	Symbol	Test Conditions	Min.	Typ.	Max.	Units
Optimized Accuracy Range	I_P		–20	–	20	A
Sensitivity	Sens	Over full range of I_P, T_A = 25°C	96	100	104	mV/A
Noise	$V_{NOISE(PP)}$	Peak-to-peak, T_A = 25°C, 100 mV/A programmed Sensitivity, C_F = 47 nF, C_{OUT} = open, 2 kHz bandwidth	–	11	–	mV
Zero Current Output Slope	$\Delta V_{OUT(Q)}$	T_A = –40°C to 25°C	–	–0.34	–	mV/°C
		T_A = 25°C to 150°C	–	–0.07	–	mV/°C
Sensitivity Slope	∆Sens	T_A = –40°C to 25°C	–	0.017	–	mV/A/°C
		T_A = 25°C to 150°C	–	–0.004	–	mV/A/°C
Total Output Error[2]	E_{TOT}	I_P =±20 A, T_A = 25°C	–	±1.5	–	%

[1] Device may be operated at higher primary current levels, I_P, and ambient temperatures, T_A, provided that the Maximum Junction Temperature, T_J(max), is not exceeded.

[2] Percentage of I_P, with I_P = 20 A. Output filtered.

x30A PERFORMANCE CHARACTERISTICS[1] T_A = –40°C to 85°C, C_F = 1 nF, and V_{CC} = 5 V, unless otherwise specified

Characteristic	Symbol	Test Conditions	Min.	Typ.	Max.	Units
Optimized Accuracy Range	I_P		–30	–	30	A
Sensitivity	Sens	Over full range of I_P, T_A = 25°C	63	66	69	mV/A
Noise	$V_{NOISE(PP)}$	Peak-to-peak, T_A = 25°C, 66 mV/A programmed Sensitivity, C_F = 47 nF, C_{OUT} = open, 2 kHz bandwidth	–	7	–	mV
Zero Current Output Slope	$\Delta V_{OUT(Q)}$	T_A = –40°C to 25°C	–	–0.35	–	mV/°C
		T_A = 25°C to 150°C	–	–0.08	–	mV/°C
Sensitivity Slope	∆Sens	T_A = –40°C to 25°C	–	0.007	–	mV/A/°C
		T_A = 25°C to 150°C	–	–0.002	–	mV/A/°C
Total Output Error[2]	E_{TOT}	I_P = ±30 A, T_A = 25°C	–	±1.5	–	%

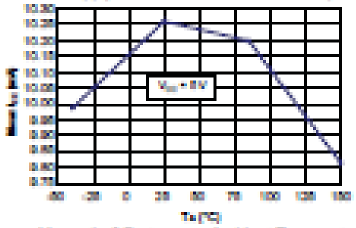
[1] Device may be operated at higher primary current levels, I_P, and ambient temperatures, T_A, provided that the Maximum Junction Temperature, T_J(max), is not exceeded.

[2] Percentage of I_P, with I_P = 30 A. Output filtered.

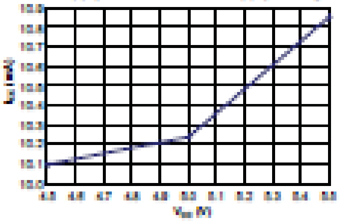
Characteristic Performance

I_P = 5 A, unless otherwise specified

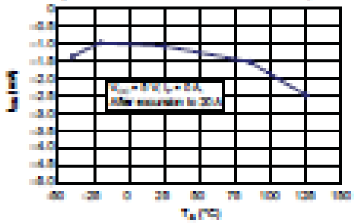
Mean Supply Current versus Ambient Temperature



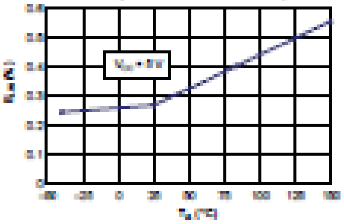
Supply Current versus Supply Voltage



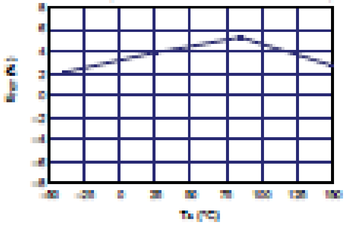
Magnetic Offset versus Ambient Temperature



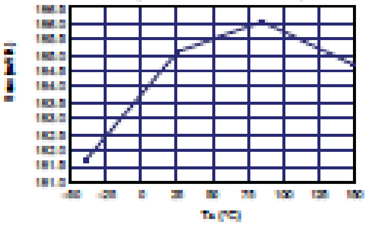
Nonlinearity versus Ambient Temperature



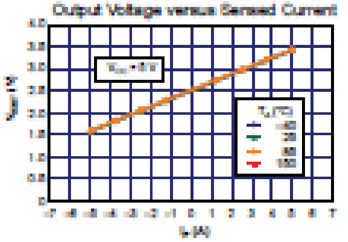
Mean Total Output Error versus Ambient Temperature



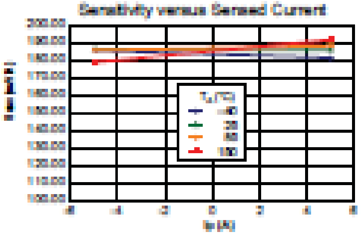
Sensitivity versus Ambient Temperature



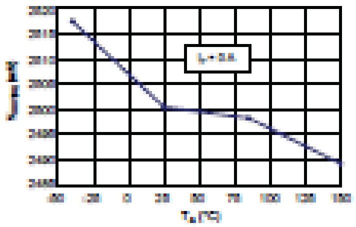
Output Voltage versus Sensed Current



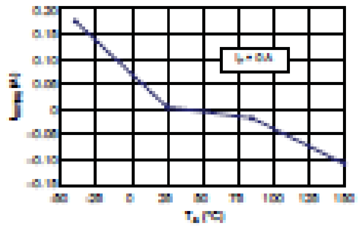
Sensitivity versus Sensed Current



0 A Output Voltage versus Ambient Temperature



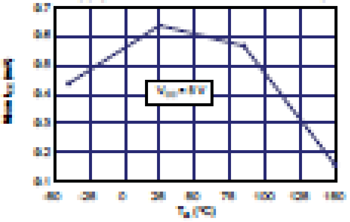
0 A Output Voltage Current versus Ambient Temperature



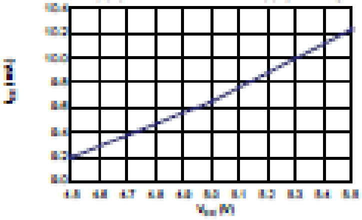
Characteristic Performance

I_P = 20 A, unless otherwise specified

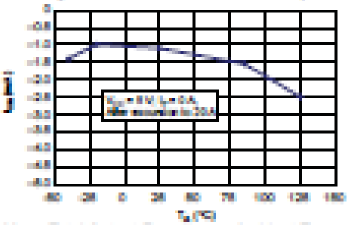
Mean Supply Current versus Ambient Temperature



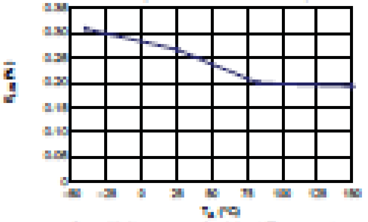
Supply Current versus Supply Voltage



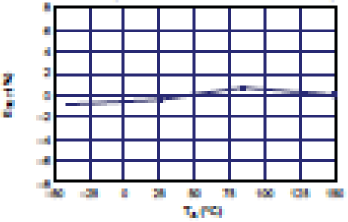
Magnetic Offset versus Ambient Temperature



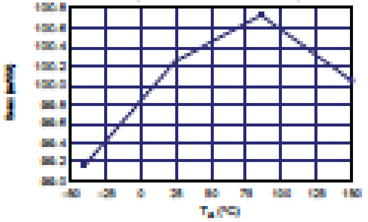
Nonlinearity versus Ambient Temperature



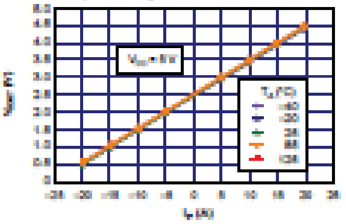
Mean Total Output Error versus Ambient Temperature



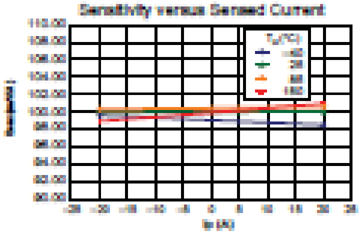
Sensitivity versus Ambient Temperature



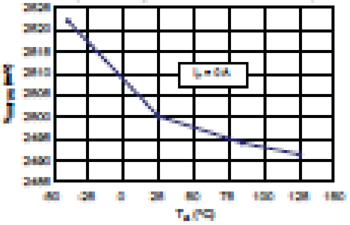
Output Voltage versus Sensed Current



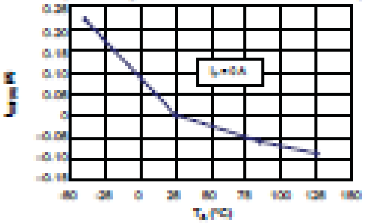
Sensitivity versus Sensed Current



0 A Output Voltage versus Ambient Temperature

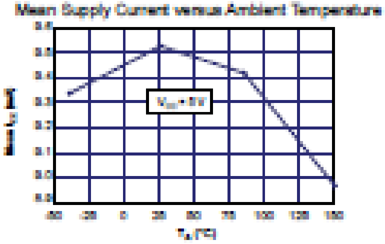


0 A Output Voltage Current versus Ambient Temperature

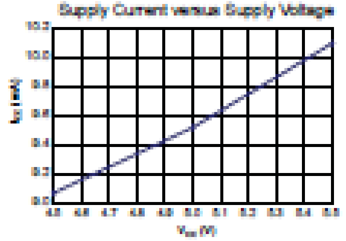


Characteristic Performance

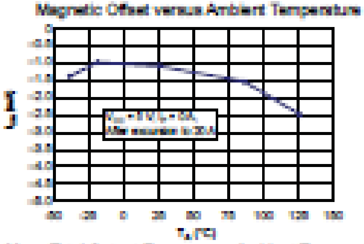
I_P = 30 A, unless otherwise specified



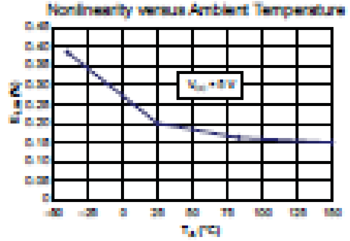
Mean Supply Current versus Ambient Temperature



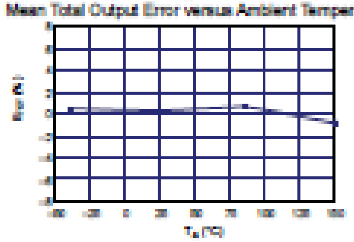
Supply Current versus Supply Voltage



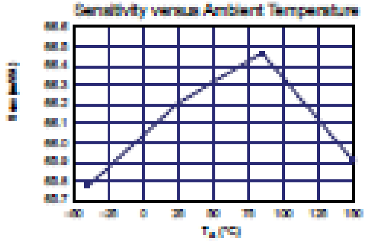
Magnetic Offset versus Ambient Temperature



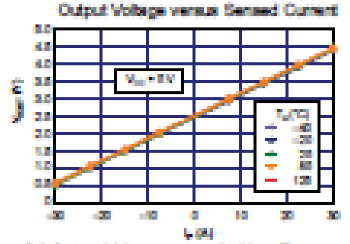
Nonlinearity versus Ambient Temperature



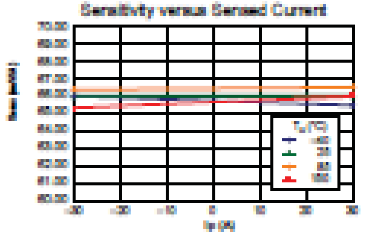
Mean Total Output Error versus Ambient Temperature



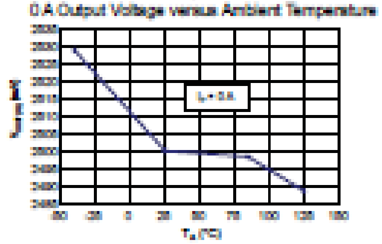
Sensitivity versus Ambient Temperature



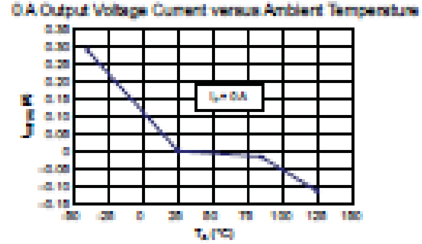
Output Voltage versus Sensed Current



Sensitivity versus Sensed Current



0 A Output Voltage versus Ambient Temperature



0 A Output Voltage Current versus Ambient Temperature

Definitions of Accuracy Characteristics

Sensitivity (Sens). The change in device output in response to a 1A change through the primary conductor. The sensitivity is the product of the magnetic circuit sensitivity (G/A) and the linear IC amplifier gain (mV/G). The linear IC amplifier gain is programmed at the factory to optimize the sensitivity (mV/A) for the full-scale current of the device.

Noise (V_{NOISE}). The product of the linear IC amplifier gain (mV/G) and the noise floor for the Allegro Hall effect linear IC (≈1 G). The noise floor is derived from the thermal and shot noise observed in Hall elements. Dividing the noise (mV) by the sensitivity (mV/A) provides the smallest current that the device is able to resolve.

Linearity (E_{LIN}). The degree to which the voltage output from the IC varies in direct proportion to the primary current through its full-scale amplitude. Nonlinearity in the output can be attributed to the saturation of the flux concentrator approaching the full-scale current. The following equation is used to derive the linearity:

$$100\left\{1-\left[\frac{\Delta\text{ gain}\times\%\text{ sat}\,(V_{IOUT\_\text{full-scale amperes}}-V_{IOUT(Q)})}{2\,(V_{IOUT\_\text{half-scale amperes}}-V_{IOUT(Q)})}\right]\right\}$$

where $V_{IOUT\_\text{full-scale amperes}}$ = the output voltage (V) when the sampled current approximates full-scale $\pm I_P$.

Symmetry (E_{SYM}). The degree to which the absolute voltage output from the IC varies in proportion to either a positive or negative full-scale primary current. The following formula is used to derive symmetry:

$$100\left(\frac{V_{IOUT\_+\text{full-scale amperes}}-V_{IOUT(Q)}}{V_{IOUT(Q)}-V_{IOUT\_-\text{full-scale amperes}}}\right)$$

Quiescent output voltage ($V_{IOUT(Q)}$). The output of the device when the primary current is zero. For a unipolar supply voltage, it nominally remains at $V_{CC}/2$. Thus, V_{CC} = 5 V translates into $V_{IOUT(Q)}$ = 2.5 V. Variation in $V_{IOUT(Q)}$ can be attributed to the resolution of the Allegro linear IC quiescent voltage trim and thermal drift.

Electrical offset voltage (V_{OE}). The deviation of the device output from its ideal quiescent value of $V_{CC}/2$ due to nonmagnetic causes. To convert this voltage to amperes, divide by the device sensitivity, Sens.

Accuracy (E_{TOT}). The accuracy represents the maximum deviation of the actual output from its ideal value. This is also known as the total output error. The accuracy is illustrated graphically in the output voltage versus current chart at right.

Accuracy is divided into four areas:

- **0 A at 25°C.** Accuracy at the zero current flow at 25°C, without the effects of temperature.
- **0 A over Δ temperature.** Accuracy at the zero current flow including temperature effects.
- **Full-scale current at 25°C.** Accuracy at the the full-scale current at 25°C, without the effects of temperature.
- **Full-scale current over Δ temperature.** Accuracy at the full-scale current flow including temperature effects.

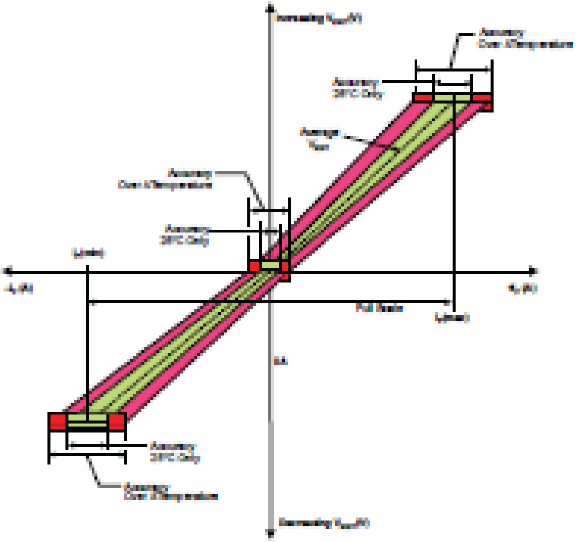
Ratiometry. The ratiometric feature means that its 0 A output, $V_{IOUT(Q)}$, (nominally equal to $V_{CC}/2$) and sensitivity, Sens, are proportional to its supply voltage, V_{CC}. The following formula is used to derive the ratiometric change in 0 A output voltage, $\Delta V_{IOUT(Q)RAT}$ (%).

$$100\left(\frac{V_{IOUT(Q)VCC}\,/\,V_{IOUT(Q)5V}}{V_{CC}\,/\,5\text{ V}}\right)$$

The ratiometric change in sensitivity, ΔSens_{RAT} (%), is defined as:

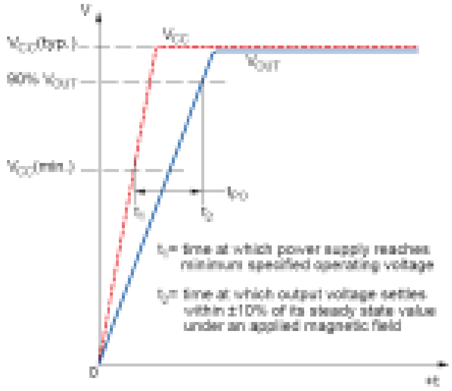
$$100\left(\frac{\text{Sens}_{VCC}\,/\,\text{Sens}_{5V}}{V_{CC}\,/\,5\text{ V}}\right)$$

Output Voltage versus Sampled Current
Accuracy at 0 A and at Full-Scale Current

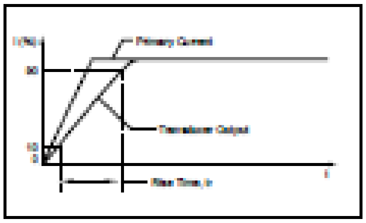


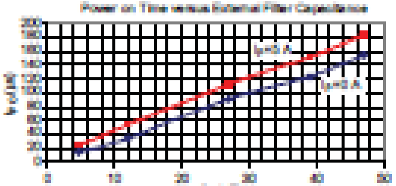
Definitions of Dynamic Response Characteristics

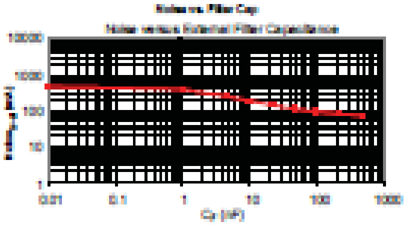
Power-On Time (t_{PO}). When the supply is ramped to its operating voltage, the device requires a finite time to power its internal components before responding to an input magnetic field. Power-On Time, t_{PO}, is defined as the time it takes for the output voltage to settle within ±10% of its steady state value under an applied magnetic field, after the power supply has reached its minimum specified operating voltage, V_{CC}(min), as shown in the chart at right.

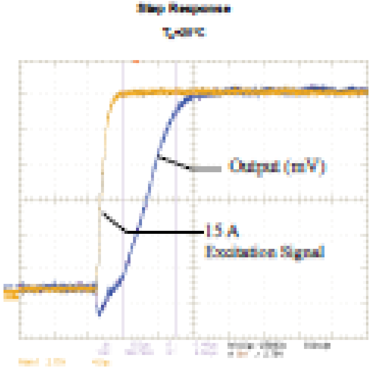


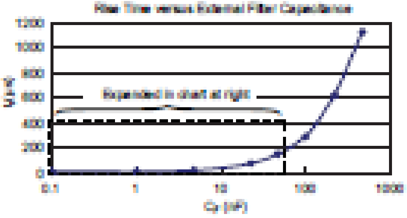
Rise time (t_r). The time interval between a) when the device reaches 10% of its full scale value, and b) when it reaches 90% of its full scale value. The rise time to a step response is used to derive the bandwidth of the device, in which $f(-3\text{ dB}) = 0.35/t_r$. Both t_r and $t_{RESPONSE}$ are detrimentally affected by eddy current losses observed in the conductive IC ground plane.



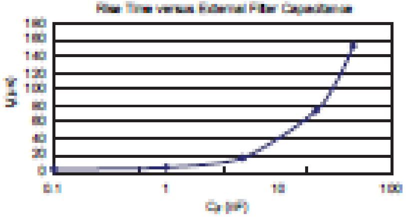








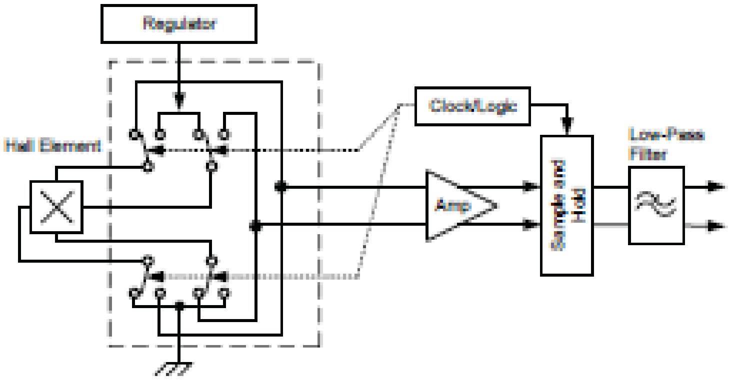
Cf (nF)	tr (µs)
Open	3.5
1	5.8
4.7	17.5
22	73.5
47	88.2
100	291.3
220	623
470	1120



Chopper Stabilization Technique

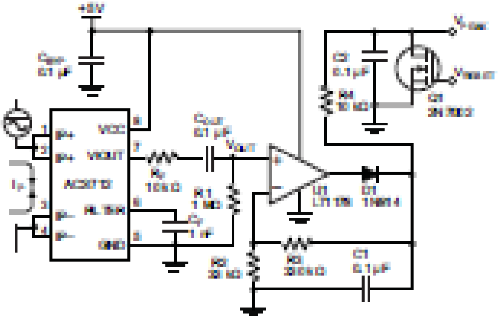
Chopper Stabilization is an innovative circuit technique that is used to minimize the offset voltage of a Hall element and an associated on-chip amplifier. Allegro patented a Chopper Stabilization technique that nearly eliminates Hall IC output drift induced by temperature or package stress effects. This offset reduction technique is based on a signal modulation-demodulation process. Modulation is used to separate the undesired DC offset signal from the magnetically induced signal in the frequency domain. Then, using a low-pass filter, the modulated DC offset is suppressed while the magnetically induced signal passes through the filter. As a result of this chopper stabilization approach, the output voltage from the Hall IC is desensitized to the effects of temperature and mechanical stress. This technique produces devices that have an extremely stable Electrical Offset Voltage, are immune to thermal stress, and have precise recoverability after temperature cycling.

This technique is made possible through the use of a BiCMOS process that allows the use of low-offset and low-noise amplifiers in combination with high-density logic integration and sample and hold circuits.

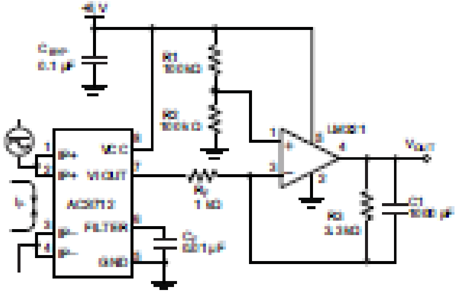


Concept of Chopper Stabilization Technique

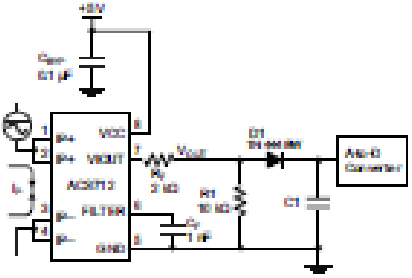
Typical Applications



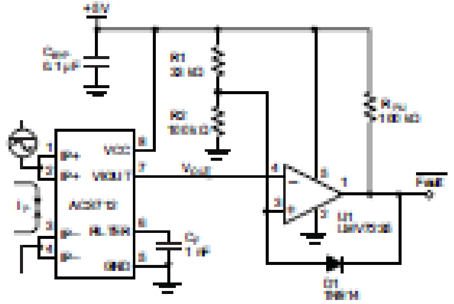
Application 2. Peak Detecting Circuit



Application 3. This configuration increases gain to 610 mV/A (tested using the ACS712ELC-05A).



Application 4. Rectified Output. 3.3 V scaling and rectification application for A-to-D converters. Replaces current transformer solutions with simpler ACS circuit. C1 is a function of the load resistance and filtering desired. R1 can be omitted if the full range is desired.



Application 5. 10 A Overcurrent Fault Latch. Fault threshold set by R1 and R2. This circuit latches an overcurrent fault and holds it until the 5 V rail is powered down.

Improving Sensing System Accuracy Using the FILTER Pin

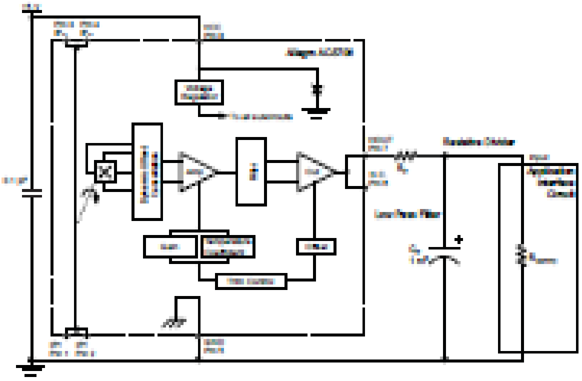
In low-frequency sensing applications, it is often advantageous to add a simple RC filter to the output of the device. Such a low-pass filter improves the signal-to-noise ratio, and therefore the resolution, of the device output signal. However, the addition of an RC filter to the output of a sensor IC can result in undesirable device output attenuation — even for DC signals.

Signal attenuation, ΔV_{ATT}, is a result of the resistive divider effect between the resistance of the external filter, R_F (see Application 6), and the input impedance and resistance of the customer interface circuit, R_{INTFC}. The transfer function of this resistive divider is given by:

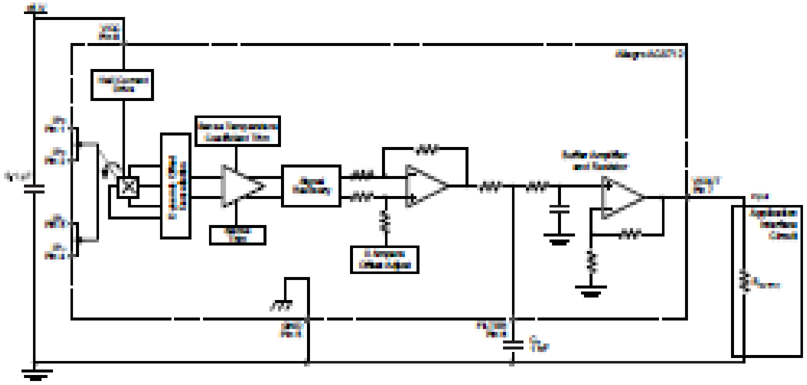
$$\Delta V_{ATT} = V_{IOUT}\left(\frac{R_{INTFC}}{R_F + R_{INTFC}}\right) .$$

Even if R_F and R_{INTFC} are designed to match, the two individual resistance values will most likely drift by different amounts over temperature. Therefore, signal attenuation will vary as a function of temperature. Note that, in many cases, the input impedance, R_{INTFC}, of a typical analog-to-digital converter (ADC) can be as low as 10 kΩ.

The ACS712 contains an internal resistor, a FILTER pin connection to the printed circuit board, and an internal buffer amplifier. With this circuit architecture, users can implement a simple RC filter via the addition of a capacitor, C_F (see Application 7) from the FILTER pin to ground. The buffer amplifier inside of the ACS712 (located after the internal resistor and FILTER pin connection) eliminates the attenuation caused by the resistive divider effect described in the equation for ΔV_{ATT}. Therefore, the ACS712 device is ideal for use in high-accuracy applications that cannot afford the signal attenuation associated with the use of an external RC low-pass filter.

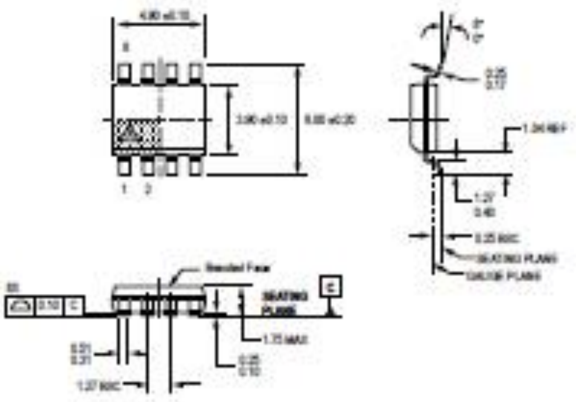


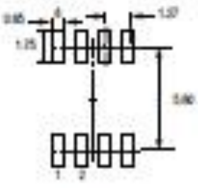
Application 6. When a low pass filter is constructed externally to a standard Hall effect device, a resistive divider may exist between the filter resistor, R_F, and the resistance of the customer interface circuit, R_{INTFC}. This resistive divider will cause excessive attenuation, as given by the transfer function for ΔV_{ATT}.



Application 7. Using the FILTER pin provided on the ACS712 eliminates the attenuation effects of the resistor divider between R_F and R_{INTFC}, shown in Application 6.

Package LC, 8-pin SOIC





PCB Layout Reference View

Standard Branding Reference View

N = Device part number
T = Device temperature range
P = Package Designator
A = Amperage
L = Lot number
Belly Brand = Country of Origin

For Reference Only; not for tooling use (reference MS-012AA)
Dimensions in millimeters
Dimensions exclusive of mold flash, gate burrs, and dambar protrusions
Exact case and lead configuration at supplier discretion within limits shown

Terminal #1 mark area

Branding scale and appearance at supplier discretion

Reference land pattern layout (reference IPC7351 SOIC127P600X175-8M); all pads a minimum of 0.20 mm from all adjacent pads; adjust as necessary to meet application process requirements and PCB layout tolerances

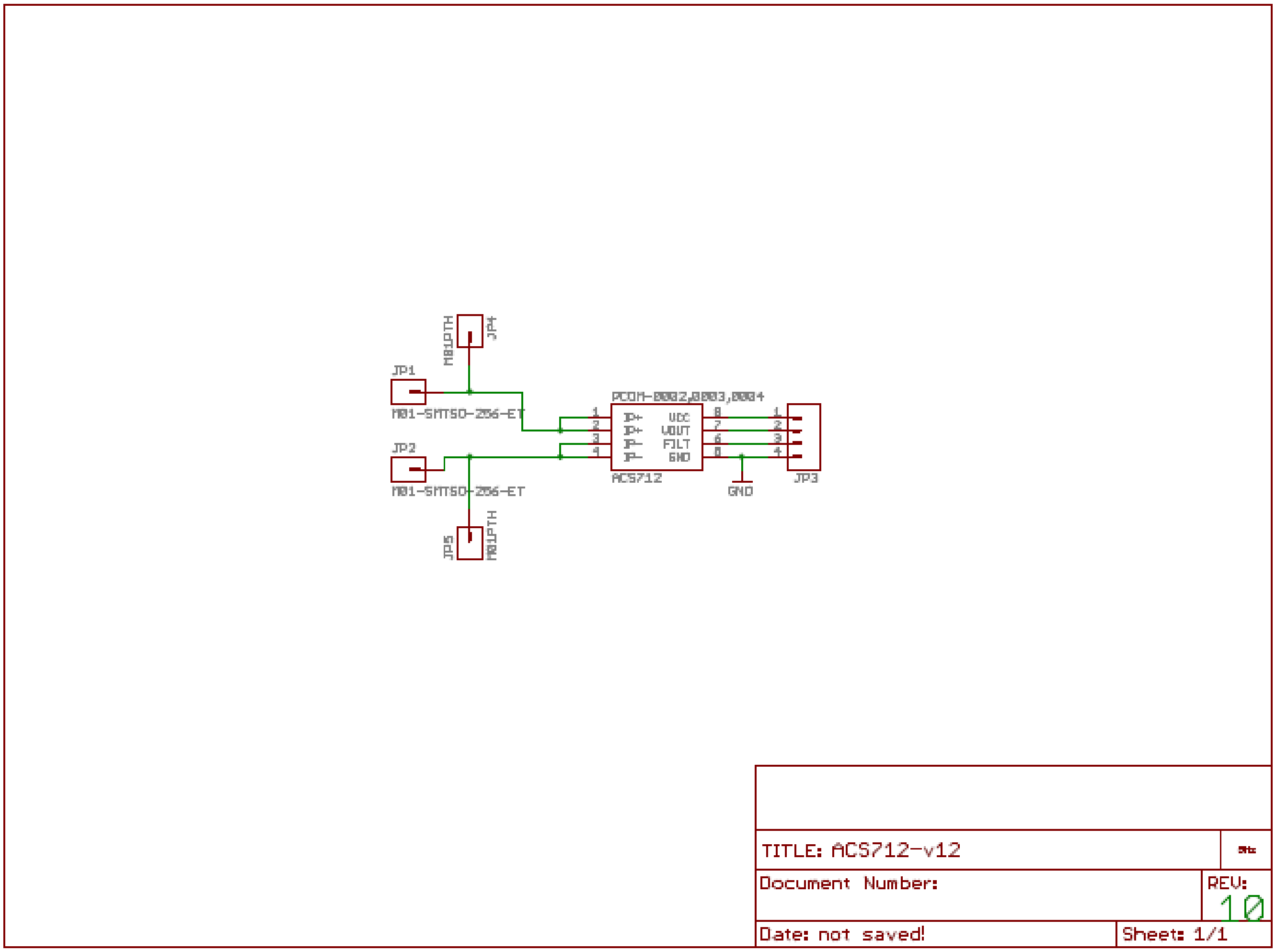
Revision History

Revision	Revision Date	Description of Revision
Rev. 15	November 16, 2012	Update rise time and isolation, I_{OUT} reference data, patents

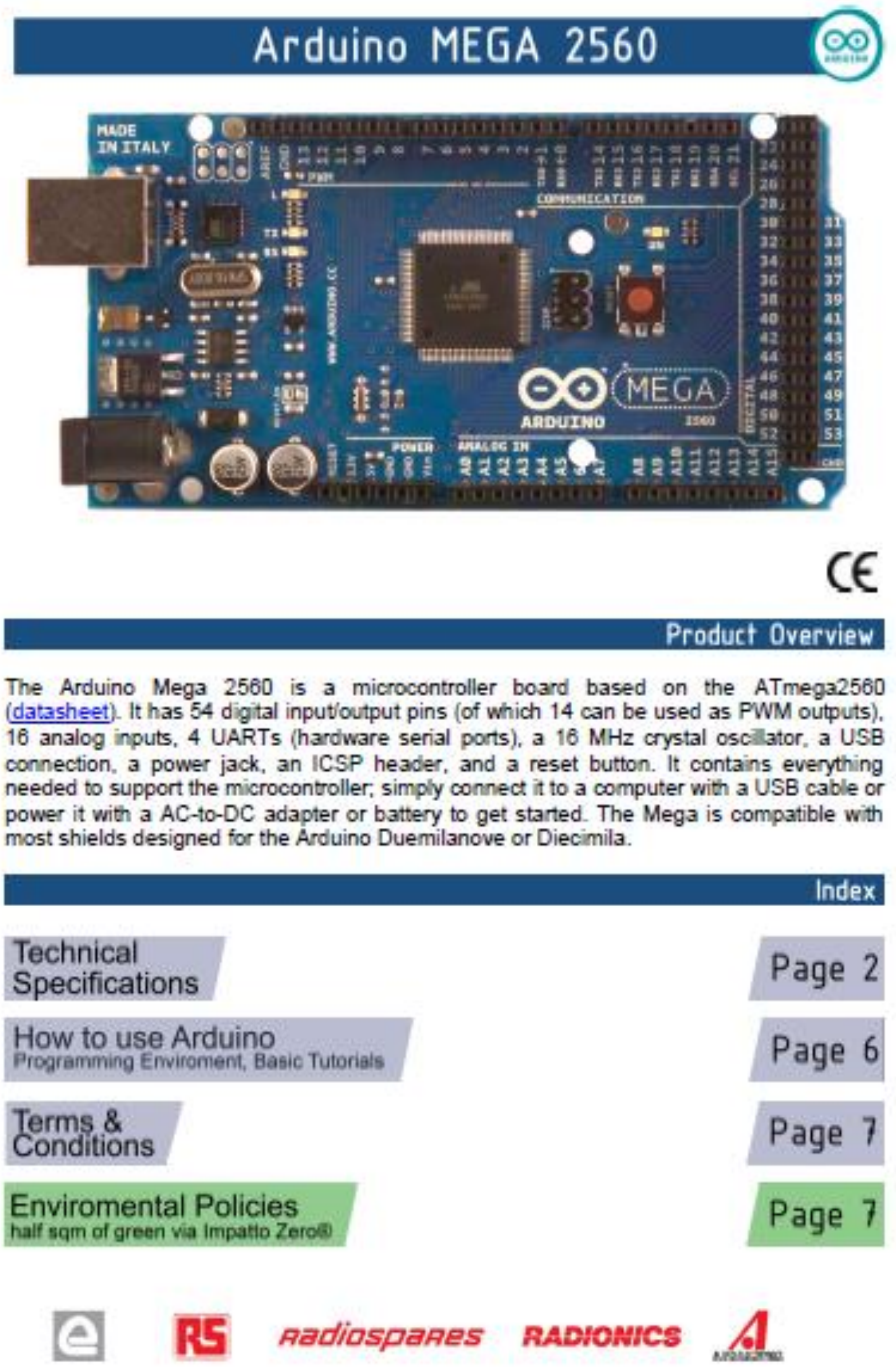
For the latest version of this document, visit our website:

www.allegromicro.com

Anexo B Diagrama Esquemático del Sensor de Corriente



Anexo C. ARDUINO MEGA 2560



Arduino MEGA 2560

Product Overview

The Arduino Mega 2560 is a microcontroller board based on the ATmega2560 (datasheet). It has 54 digital input/output pins (of which 14 can be used as PWM outputs), 16 analog inputs, 4 UARTs (hardware serial ports), a 16 MHz crystal oscillator, a USB connection, a power jack, an ICSP header, and a reset button. It contains everything needed to support the microcontroller; simply connect it to a computer with a USB cable or power it with a AC-to-DC adapter or battery to get started. The Mega is compatible with most shields designed for the Arduino Duemilanove or Diecimila.

Index

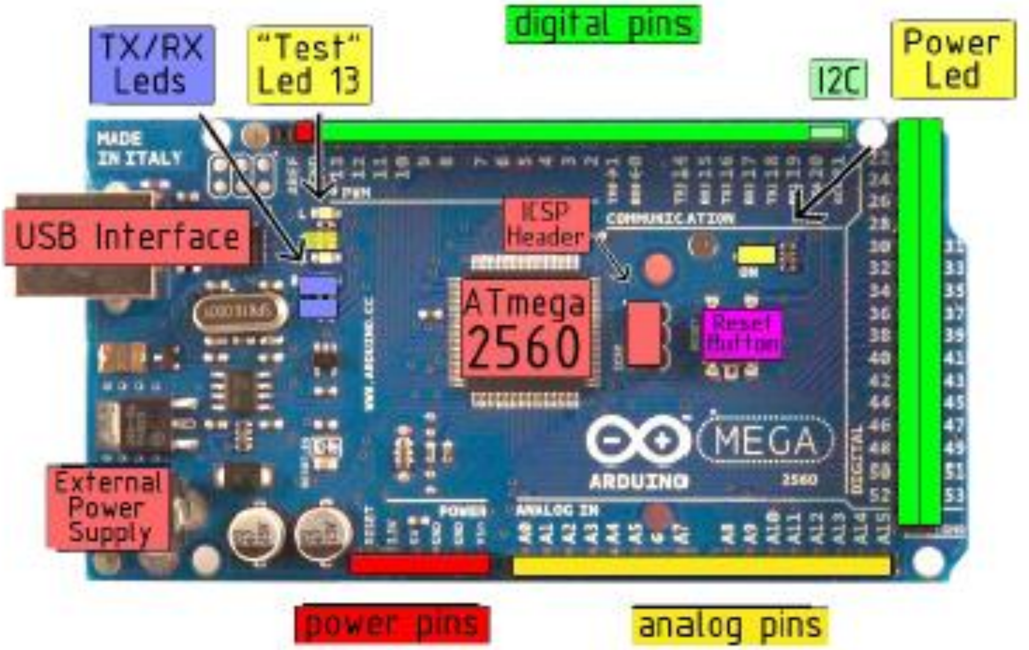
Technical Specification

EAGLE files: arduino-mega2560-reference-design.zip Schematic: arduino-mega2560-schematic.pdf

Summary

Microcontroller	ATmega2560
Operating Voltage	5V
Input Voltage (recommended)	7-12V
Input Voltage (limits)	6-20V
Digital I/O Pins	54 (of which 14 provide PWM output)
Analog Input Pins	16
DC Current per I/O Pin	40 mA
DC Current for 3.3V Pin	50 mA
Flash Memory	256 KB of which 8 KB used by bootloader
SRAM	8 KB
EEPROM	4 KB
Clock Speed	16 MHz

the board



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Power

The Arduino Mega2560 can be powered via the USB connection or with an external power supply. The power source is selected automatically. External (non-USB) power can come either from an AC-to-DC adapter (wall-wart) or battery. The adapter can be connected by plugging a 2.1mm center-positive plug into the board's power jack. Leads from a battery can be inserted in the Gnd and Vin pin headers of the POWER connector.

The board can operate on an external supply of 6 to 20 volts. If supplied with less than 7V, however, the 5V pin may supply less than five volts and the board may be unstable. If using more than 12V, the voltage regulator may overheat and damage the board. The recommended range is 7 to 12 volts.

The Mega2560 differs from all preceding boards in that it does not use the FTDI USB-to-serial driver chip. Instead, it features the Atmega8U2 programmed as a USB-to-serial converter.

The power pins are as follows:

- **VIN.** The input voltage to the Arduino board when it's using an external power source (as opposed to 5 volts from the USB connection or other regulated power source). You can supply voltage through this pin, or, if supplying voltage via the power jack, access it through this pin.
- **5V.** The regulated power supply used to power the microcontroller and other components on the board. This can come either from VIN via an on-board regulator, or be supplied by USB or another regulated 5V supply.
- **3V3.** A 3.3 volt supply generated by the on-board regulator. Maximum current draw is 50 mA.
- **GND.** Ground pins.

Memory

The ATmega2560 has 256 KB of flash memory for storing code (of which 8 KB is used for the bootloader), 8 KB of SRAM and 4 KB of EEPROM (which can be read and written with the EEPROM library).

Input and Output

Each of the 54 digital pins on the Mega can be used as an input or output, using pinMode(), digitalWrite(), and digitalRead() functions. They operate at 5 volts. Each pin can provide or receive a maximum of 40 mA and has an internal pull-up resistor (disconnected by default) of 20-50 kOhms. In addition, some pins have specialized functions:

- **Serial: 0 (RX) and 1 (TX); Serial 1: 19 (RX) and 18 (TX); Serial 2: 17 (RX) and 16 (TX); Serial 3: 15 (RX) and 14 (TX).** Used to receive (RX) and transmit (TX) TTL serial data. Pins 0 and 1 are also connected to the corresponding pins of the ATmega8U2 USB-to-TTL Serial chip .
- **External Interrupts: 2 (interrupt 0), 3 (interrupt 1), 18 (interrupt 5), 19 (interrupt 4), 20 (interrupt 3), and 21 (interrupt 2).** These pins can be configured to trigger an interrupt on a low value, a rising or falling edge, or a change in value. See the attachInterrupt() function for details.
- **PWM: 0 to 13.** Provide 8-bit PWM output with the analogWrite() function.
- **SPI: 50 (MISO), 51 (MOSI), 52 (SCK), 53 (SS).** These pins support SPI communication, which, although provided by the underlying hardware, is not currently included in the Arduino language. The SPI pins are also broken out on the ICSP header, which is physically compatible with the Duemilanove and Diecimila.
- **LED: 13.** There is a built-in LED connected to digital pin 13. When the pin is HIGH value, the LED is on, when the pin is LOW, it's off.
- **I^2C: 20 (SDA) and 21 (SCL).** Support I^2C (TWI) communication using the Wire library (documentation on the Wiring website). Note that these pins are not in the same location as the I^2C pins on the Duemilanove.

The Mega2560 has 16 analog inputs, each of which provide 10 bits of resolution (i.e. 1024 different values). By default they measure from ground to 5 volts, though is it possible to change the upper end of their range using the AREF pin and analogReference() function.

There are a couple of other pins on the board:

- **AREF.** Reference voltage for the analog inputs. Used with analogReference().
- **Reset.** Bring this line LOW to reset the microcontroller. Typically used to add a reset button to shields which block the one on the board.

Communication

The Arduino Mega2560 has a number of facilities for communicating with a computer, another Arduino, or other microcontrollers. The ATmega2560 provides four hardware UARTs for TTL (5V) serial communication. An ATmega8U2 on the board channels one of these over USB and provides a virtual com port to software on the computer (Windows machines will need a .inf file, but OSX and Linux machines will recognize the board as a COM port automatically. The Arduino software includes a serial monitor which allows simple textual data to be sent to and from the board. The RX and TX LEDs on the board will flash when data is being transmitted via the ATmega8U2 chip and USB connection to the computer (but not for serial communication on pins 0 and 1).

A SoftwareSerial library allows for serial communication on any of the Mega's digital pins.

The ATmega2560 also supports I2C (TWI) and SPI communication. The Arduino software includes a Wire library to simplify use of the I2C bus; see the documentation on the Wiring website for details. To use the SPI communication, please see the ATmega2560 datasheet.

Programming

The Arduino Mega2560 can be programmed with the Arduino software (download). For details, see the reference and tutorials.

The Atmega2560 on the Arduino Mega comes preburned with a bootloader that allows you to upload new code to it without the use of an external hardware programmer. It communicates using the original STK500 protocol (reference, C header files).

You can also bypass the bootloader and program the microcontroller through the ICSP (In-Circuit Serial Programming) header; see these instructions for details.

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Automatic (Software) Reset

Rather then requiring a physical press of the reset button before an upload, the Arduino Mega2560 is designed in a way that allows it to be reset by software running on a connected computer. One of the hardware flow control lines (DTR) of the ATmega8U2 is connected to the reset line of the ATmega2560 via a 100 nanofarad capacitor. When this line is asserted (taken low), the reset line drops long enough to reset the chip. The Arduino software uses this capability to allow you to upload code by simply pressing the upload button in the Arduino environment. This means that the bootloader can have a shorter timeout, as the lowering of DTR can be well-coordinated with the start of the upload.

This setup has other implications. When the Mega2560 is connected to either a computer running Mac OS X or Linux, it resets each time a connection is made to it from software (via USB). For the following half-second or so, the bootloader is running on the Mega2560. While it is programmed to ignore malformed data (i.e. anything besides an upload of new code), it will intercept the first few bytes of data sent to the board after a connection is opened. If a sketch running on the board receives one-time configuration or other data when it first starts, make sure that the software with which it communicates waits a second after opening the connection and before sending this data.

The Mega contains a trace that can be cut to disable the auto-reset. The pads on either side of the trace can be soldered together to re-enable it. It's labeled "RESET-EN". You may also be able to disable the auto-reset by connecting a 110 ohm resistor from 5V to the reset line; see this forum thread for details.

USB Overcurrent Protection

The Arduino Mega has a resettable polyfuse that protects your computer's USB ports from shorts and overcurrent. Although most computers provide their own internal protection, the fuse provides an extra layer of protection. If more than 500 mA is applied to the USB port, the fuse will automatically break the connection until the short or overload is removed.

Physical Characteristics and Shield Compatibility

The maximum length and width of the Mega PCB are 4 and 2.1 inches respectively, with the USB connector and power jack extending beyond the former dimension. Three screw holes allow the board to be attached to a surface or case. Note that the distance between digital pins 7 and 8 is 160 mil (0.16"), not an even multiple of the 100 mil spacing of the other pins.

The Mega is designed to be compatible with most shields designed for the Diecimila or Duemilanove. Digital pins 0 to 13 (and the adjacent AREF and GND pins), analog inputs 0 to 5, the power header, and ICSP header are all in equivalent locations. Further the main UART (serial port) is located on the same pins (0 and 1), as are external interrupts 0 and 1 (pins 2 and 3 respectively). SPI is available through the ICSP header on both the Mega and Duemilanove / Diecimila. **Please note that I²C is not located on the same pins on the Mega (20 and 21) as the Duemilanove / Diecimila (analog inputs 4 and 5).**

How to use Arduino

Arduino can sense the environment by receiving input from a variety of sensors and can affect its surroundings by controlling lights, motors, and other actuators. The microcontroller on the board is programmed using the [Arduino programming language]() (based on [Wiring]()) and the Arduino development environment (based on [Processing]()). Arduino projects can be stand-alone or they can communicate with software on running on a computer (e.g. Flash, Processing, MaxMSP).

Arduino is a cross-platoform program. You'll have to follow different instructions for your personal OS. Check on the [Arduino site]() for the latest instructions. *http://arduino.cc/en/Guide/HomePage*

Linux Install **Windows Install** **Mac Install**

Once you have downloaded/unzipped the arduino IDE, you can Plug the Arduino to your PC via USB cable.

Blink led

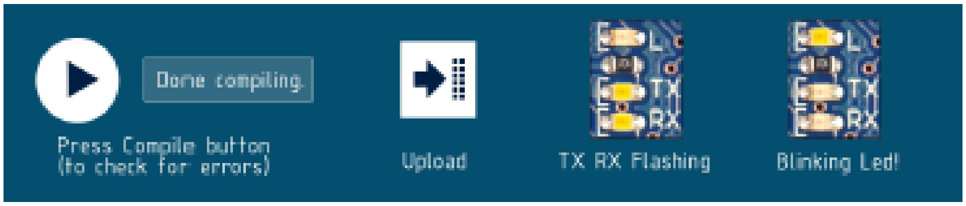
Now you're actually ready to "burn" your first program on the arduino board. To select "blink led", the physical translation of the well known programming "hello world", select

File>Sketchbook>
Arduino-0017>Examples>
Digital>Blink

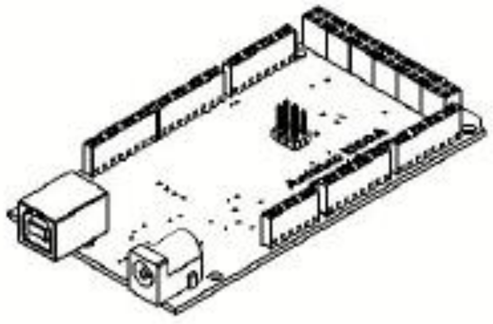
Once you have your skecth you'll see something very close to the screenshot on the right.

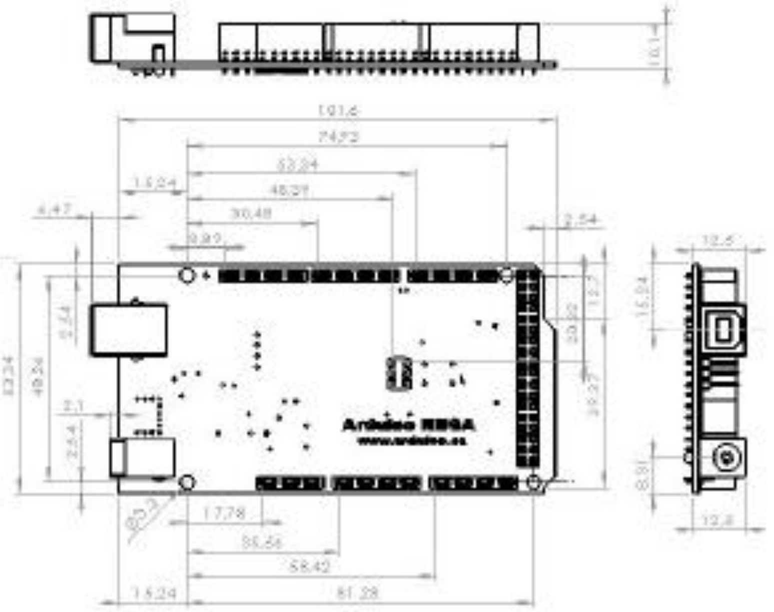
In **Tools>Board** select MEGA

Now you have to go to
Tools>SerialPort
and select the right serial port, the one arduino is attached to.



Dimensioned Drawing





Terms & Conditions

1. Warranties

1.1 The producer warrants that its products will conform to the Specifications. This warranty lasts for one (1) years from the date of the sale. The producer shall not be liable for any defects that are caused by neglect, misuse or mistreatment by the Customer, including improper installation or testing, or for any products that have been altered or modified in any way by a Customer. Moreover, The producer shall not be liable for any defects that result from Customer's design, specifications or instructions for such products. Testing and other quality control techniques are used to the extent the producer deems necessary.

1.2 If any products fail to conform to the warranty set forth above, the producer's sole liability shall be to replace such products. The producer's liability shall be limited to products that are determined by the producer not to conform to such warranty. If the producer elects to replace such products, the producer shall have a reasonable time to replacements. Replaced products shall be warranted for a new full warranty period.

1.3 EXCEPT AS SET FORTH ABOVE, PRODUCTS ARE PROVIDED "AS IS" AND "WITH ALL FAULTS." THE PRODUCER DISCLAIMS ALL OTHER WARRANTIES, EXPRESS OR IMPLIED, REGARDING PRODUCTS, INCLUDING BUT NOT LIMITED TO, ANY IMPLIED WARRANTIES OF MERCHANTABILITY OR FITNESS FOR A PARTICULAR PURPOSE

1.4 Customer agrees that prior to using any systems that include the producer products, Customer will test such systems and the functionality of the products as used in such systems. The producer may provide technical, applications or design advice, quality characterization, reliability data or other services. Customer acknowledges and agrees that providing these services shall not expand or otherwise alter the producer's warranties, as set forth above, and no additional obligations or liabilities shall arise from the producer providing such services.

1.5 The Arduino™ products are not authorized for use in safety-critical applications where a failure of the product would reasonably be expected to cause severe personal injury or death. Safety-Critical Applications include, without limitation, life support devices and systems, equipment or systems for the operation of nuclear facilities and weapons systems. Arduino™ products are neither designed nor intended for use in military or aerospace applications or environments and for automotive applications or environment. Customer acknowledges and agrees that any such use of Arduino™ products which is solely at the Customer's risk, and that Customer is solely responsible for compliance with all legal and regulatory requirements in connection with such use.

1.6 Customer acknowledges and agrees that it is solely responsible for compliance with all legal, regulatory and safety-related requirements concerning its products and any use of Arduino™ products in Customer's applications, notwithstanding any applications-related information or support that may be provided by the producer.

2. Indemnification

The Customer acknowledges and agrees to defend, indemnify and hold harmless the producer from and against any and all third-party losses, damages, liabilities and expenses it incurs to the extent directly caused by: (i) an actual breach by a Customer of the representation and warranties made under this terms and conditions or (ii) the gross negligence or willful misconduct by the Customer.

3. Consequential Damages Waiver

In no event the producer shall be liable to the Customer or any third parties for any special, collateral, indirect, punitive, incidental, consequential or exemplary damages in connection with or arising out of the products provided hereunder, regardless of whether the producer has been advised of the possibility of such damages. This section will survive the termination of the warranty period.

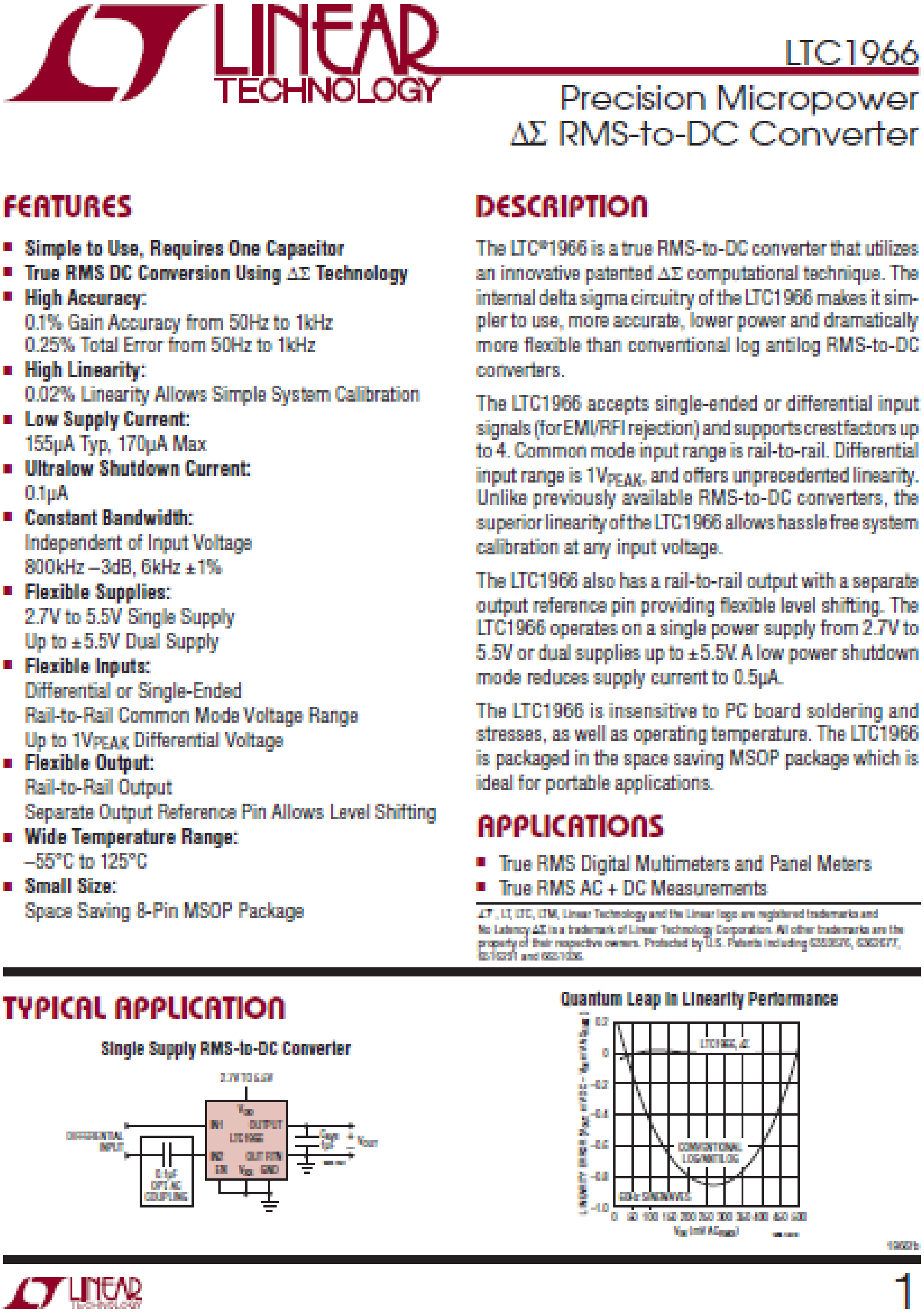
4. Changes to specifications

The producer may make changes to specifications and product descriptions at any time, without notice. The Customer must not rely on the absence or characteristics of any features or instructions marked "reserved" or "undefined." The producer reserves these for future definition and shall have no responsibility whatsoever for conflicts or incompatibilities arising from future changes to them. The product information on the Web Site or Materials is subject to change without notice. Do not finalize a design with this information.

Enviromental Policies

The producer of Arduino™ has joined the Impatto Zero® policy of LifeGate.it. For each Arduino board produced is created / looked after half squared Km of Costa Rica's forest's.

Anexo D. Sensor de Voltaje RMS



LTC1966

Precision Micropower ΔΣ RMS-to-DC Converter

FEATURES

- **Simple to Use, Requires One Capacitor**
- **True RMS DC Conversion Using ΔΣ Technology**
- **High Accuracy:**
  0.1% Gain Accuracy from 50Hz to 1kHz
  0.25% Total Error from 50Hz to 1kHz
- **High Linearity:**
  0.02% Linearity Allows Simple System Calibration
- **Low Supply Current:**
  155µA Typ, 170µA Max
- **Ultralow Shutdown Current:**
  0.1µA
- **Constant Bandwidth:**
  Independent of Input Voltage
  800kHz –3dB, 6kHz ±1%
- **Flexible Supplies:**
  2.7V to 5.5V Single Supply
  Up to ±5.5V Dual Supply
- **Flexible Inputs:**
  Differential or Single-Ended
  Rail-to-Rail Common Mode Voltage Range
  Up to $1V_{PEAK}$ Differential Voltage
- **Flexible Output:**
  Rail-to-Rail Output
  Separate Output Reference Pin Allows Level Shifting
- **Wide Temperature Range:**
  –55°C to 125°C
- **Small Size:**
  Space Saving 8-Pin MSOP Package

DESCRIPTION

The LTC®1966 is a true RMS-to-DC converter that utilizes an innovative patented ΔΣ computational technique. The internal delta sigma circuitry of the LTC1966 makes it simpler to use, more accurate, lower power and dramatically more flexible than conventional log antilog RMS-to-DC converters.

The LTC1966 accepts single-ended or differential input signals (for EMI/RFI rejection) and supports crest factors up to 4. Common mode input range is rail-to-rail. Differential input range is $1V_{PEAK}$, and offers unprecedented linearity. Unlike previously available RMS-to-DC converters, the superior linearity of the LTC1966 allows hassle free system calibration at any input voltage.

The LTC1966 also has a rail-to-rail output with a separate output reference pin providing flexible level shifting. The LTC1966 operates on a single power supply from 2.7V to 5.5V or dual supplies up to ±5.5V. A low power shutdown mode reduces supply current to 0.5µA.

The LTC1966 is insensitive to PC board soldering and stresses, as well as operating temperature. The LTC1966 is packaged in the space saving MSOP package which is ideal for portable applications.

APPLICATIONS

- True RMS Digital Multimeters and Panel Meters
- True RMS AC + DC Measurements

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TYPICAL APPLICATION

Single Supply RMS-to-DC Converter

Quantum Leap In Linearity Performance

1966fb

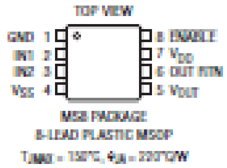
1

ABSOLUTE MAXIMUM RATINGS

(Note 1)

Supply Voltage
  V_{DD} to GND .... –0.3V to 7V
  V_{DD} to V_{SS} .... –0.3V to 12V
  V_{SS} to GND .... –7V to 0.3V
Input Currents (Note 2) .... ±10mA
Output Current (Note 3) .... ±10mA
$\overline{\text{ENABLE}}$ Voltage .... V_{SS} – 0.3V to V_{SS} + 12V
OUT RTN Voltage .... V_{SS} – 0.3V to V_{DD}
Operating Temperature Range (Note 4)
  LTC1966C/LTC1966I .... –40°C to 85°C
  LTC1966H .... –40°C to 125°C
  LTC1966MP .... –55°C to 125°C
Specified Temperature Range (Note 5)
  LTC1966C/LTC1966I .... –40°C to 85°C
  LTC1966H .... –40°C to 125°C
  LTC1966MP .... –55°C to 125°C
Maximum Junction Temperature .... 150°C
Storage Temperature Range .... –65°C to 150°C
Lead Temperature (Soldering, 10 sec)................... 300°C

PIN CONFIGURATION



ORDER INFORMATION

LEAD FREE FINISH	TAPE AND REEL	PART MARKING*	PACKAGE DESCRIPTION	TEMPERATURE RANGE
LTC1966CMS8#PBF	LTC1966CMS8#TRPBF	LTTG	8-Lead Plastic MSOP	0°C to 70°C
LTC1966IMS8#PBF	LTC1966IMS8#TRPBF	LTTH	8-Lead Plastic MSOP	–40°C to 85°C
LTC1966HMS8#PBF	LTC1966HMS8#TRPBF	LTTG	8-Lead Plastic MSOP	–40°C to 125°C
LTC1966MPMS8#PBF	LTC1966MPMS8#TRPBF	LTTG	8-Lead Plastic MSOP	–55°C to 125°C

Consult LTC Marketing for parts specified with wider operating temperature ranges. *The temperature grade is identified by a label on the shipping container.

For more information on lead free part marking, go to: http://www.linear.com/leadfree/
For more information on tape and reel specifications, go to: http://www.linear.com/tapeandreel/

ELECTRICAL CHARACTERISTICS

The ● denotes the specifications which apply over the full operating temperature range, otherwise specifications are at T_A = 25°C. V_{DD} = 5V, V_{SS} = –5V, V_{OUTRTN} = 0V, C_{AVE} = 10µF, V_{IN} = 200mV$_{RMS}$, $V_{\overline{ENABLE}}$ = 0.5V unless otherwise noted.

SYMBOL	PARAMETER	CONDITIONS		MIN	TYP	MAX	UNITS
Conversion Accuracy							
G_{ERR}	Conversion Gain Error	50Hz to 1kHz Input (Notes 6, 7) LTC1966C, LTC1966I LTC1966H, LTC1966MP	 ● ●		±0.1	±0.3 ±0.4 ±0.7	% % %
V_{OOS}	Output Offset Voltage	(Notes 6, 7) LTC1966C, LTC1966I LTC1966H, LTC1966MP	 ● ●		0.1	0.2 0.4 0.6	mV mV mV
LIN_{ERR}	Linearity Error	50mV to 350mV (Notes 7, 8)	●		0.02	0.15	%

ELECTRICAL CHARACTERISTICS

The ● denotes the specifications which apply over the full operating temperature range, otherwise specifications are at $T_A = 25°C$. $V_{DD} = 5V$, $V_{SS} = –5V$, $V_{OUTRTN} = 0V$, $C_{AVE} = 10µF$, $V_{IN} = 200mV_{RMS}$, $V_{ENABLE} = 0.5V$ unless otherwise noted.

SYMBOL	PARAMETER	CONDITIONS		MIN	TYP	MAX	UNITS
PSRR	Power Supply Rejection	(Note 9) LTC1966C, LTC1966I LTC1966H, LTC1966MP	 ● ●		0.02	0.15 0.20 0.3	%/V %/V %/V
V_{IOS}	Input Offset Voltage	(Notes 6, 7, 10)	 ●		0.02	0.8 1.0	mV mV
Accuracy vs Crest Factor (CF)							
	CF = 4	60Hz Fundamental, 200mV_{RMS} (Note 11)	●	–1		2	mV
	CF = 5	60Hz Fundamental, 200mV_{RMS} (Note 11)	●	–20		30	mV
Input Characteristics							
I_{VR}	Input Voltage Range	(Note 14)	●	V_{SS}		V_{DD}	V
Z_{IN}	Input Impedance	Average, Differential (Note 12) Average, Common Mode (Note 12)			8 100		MΩ MΩ
CMRRI	Input Common Mode Rejection	(Note 13)	●		7	200	µV/V
V_{IMAX}	Maximum Input Swing	Accuracy = 1% (Note 14)	●	1	1.05		V
V_{IMIN}	Minimum RMS Input		●			5	mV
PSRRI	Power Supply Rejection	V_{DD} Supply (Note 9) V_{SS} Supply (Note 9)	● ●		250 120	600 300	µV/V µV/V
Output Characteristics							
OVR	Output Voltage Range		●	V_{SS}		V_{DD}	V
Z_{OUT}	Output Impedance	V_{ENABLE} = 0.5V (Note 12) V_{ENABLE} = 4.5V	●	75	85 30	95	kΩ kΩ
CMRRO	Output Common Mode Rejection	(Note 13)	●		16	200	µV/V
V_{OMAX}	Maximum Differential Output Swing	Accuracy = 2%, DC Input (Note 14)	 ●	1.0 0.9	1.05		V V
PSRRO	Power Supply Rejection	V_{DD} Supply (Note 9) V_{SS} Supply (Note 9)	● ●		250 50	1000 500	µV/V µV/V
Frequency Response							
f_{1P}	1% Additional Error (Note 15)	C_{AVE} = 10µF			6		kHz
f_{10P}	10% Additional Error (Note 15)	C_{AVE} = 10µF			20		kHz
$f_{–3dB}$	±3dB Frequency (Note 15)				800		kHz
Power Supplies							
V_{DD}	Positive Supply Voltage		●	2.7		5.5	V
V_{SS}	Negative Supply Voltage	(Note 16)	●	–5.5		0	V
I_{DD}	Positive Supply Current	IN1 = 20mV, IN2 = 0V IN1 = 200mV, IN2 = 0V	●		155 158	170	µA µA
I_{SS}	Negative Supply Current	IN1 = 20mV, IN2 = 0V	●		12	20	µA
Shutdown Characteristics							
I_{DDS}	Supply Currents	V_{ENABLE} = 4.5V	●		0.5	10	µA
I_{SSS}	Supply Currents	V_{ENABLE} = 4.5V LTC1966H, LTC1966MP	● ●	–1 –2	–0.1		µA µA
I_{IH}	ENABLE Pin Current High	V_{ENABLE} = 4.5V	●	–0.3	–0.05		µA

ELECTRICAL CHARACTERISTICS

The ● denotes the specifications which apply over the full operating temperature range, otherwise specifications are at $T_A = 25°C$. $V_{DD} = 5V$, $V_{SS} = -5V$, $V_{OUTRTN} = 0V$, $C_{AVE} = 10µF$, $V_{IN} = 200mV_{RMS}$, $V_{ENABLE} = 0.5V$ unless otherwise noted.

SYMBOL	PARAMETER	CONDITIONS		MIN	TYP	MAX	UNITS
I_{IL}	ENABLE Pin Current Low	$V_{ENABLE} = 0.5V$ LTC1966H, LTC1966MP	● ●	−2 −10	−1	−0.1	µA µA
V_{TH}	ENABLE Threshold Voltage	$V_{DD} = 5V$, $V_{SS} = -5V$ $V_{DD} = 5V$, $V_{SS} = GND$ $V_{DD} = 2.7V$, $V_{SS} = GND$			2.4 2.1 1.3		V V V
V_{HYS}	ENABLE Threshold Hysteresis				0.1		V

Note 1: Stresses beyond those listed under Absolute Maximum Ratings may cause permanent damage to the device. Exposure to any Absolute Maximum Rating condition for extended periods may affect device reliability and lifetime.

Note 2: The inputs (IN1, IN2) are protected by shunt diodes to V_{SS} and V_{DD}. If the inputs are driven beyond the rails, the current should be limited to less than 10mA.

Note 3: The LTC1966 output (V_{OUT}) is high impedance and can be overdriven, either sinking or sourcing current, to the limits stated.

Note 4: The LTC1966C/LTC1966I are guaranteed functional over the operating temperature range of −40°C to 85°C. The LTC1966H/LTC1966MP are guaranteed functional over the operating temperature range of −55°C to 125°C.

Note 5: The LTC1966C is guaranteed to meet specified performance from 0°C to 70°C. The LTC1966C is designed, characterized and expected to meet specified performance from −40°C to 85°C but is not tested nor QA sampled at these temperatures. The LTC1966I is guaranteed to meet specified performance from −40°C to 85°C. The LTC1966H is guaranteed to meet specified performance from −40°C to 125°C. The LTC1966MP is guaranteed to meet specified performance from −55°C to 125°C.

Note 6: High speed automatic testing cannot be performed with $C_{AVE} = 10µF$. The LTC1966 is 100% tested with $C_{AVE} = 22nF$. Correlation tests have shown that the performance limits above can be guaranteed with the additional testing being performed to guarantee proper operation of all the internal circuitry.

Note 7: High speed automatic testing cannot be performed with 60Hz inputs. The LTC1966 is 100% tested with DC and 10kHz input signals. Measurements with DC inputs from 50mV to 350mV are used to calculate the four parameters: G_{ERR}, V_{OOS}, V_{IOS} and linearity error. Correlation tests have shown that the performance limits above can be guaranteed with the additional testing being performed to guarantee proper operation of all internal circuitry.

Note 8: The LTC1966 is inherently very linear. Unlike older log/antilog circuits, its behavior is the same with DC and AC inputs, and DC inputs are used for high speed testing.

Note 9: The power supply rejections of the LTC1966 are measured with DC inputs from 50mV to 350mV. The change in accuracy from $V_{DD} = 2.7V$ to $V_{DD} = 5.5V$ with $V_{SS} = 0V$ is divided by 2.8V. The change in accuracy from $V_{SS} = 0V$ to $V_{SS} = -5.5V$ with $V_{DD} = 5.5V$ is divided by 5.5V.

Note 10: Previous generation RMS-to-DC converters required nonlinear input stages as well as a nonlinear core. Some parts specify a DC reversal error, combining the effects of input nonlinearity and input offset voltage. The LTC1966 behavior is simpler to characterize and the input offset voltage is the only significant source of DC reversal error.

Note 11: High speed automatic testing cannot be performed with 60Hz inputs. The LTC1966 is 100% tested with DC stimulus. Correlation tests have shown that the performance limits above can be guaranteed with the additional testing being performed to verify proper operation of all internal circuitry.

Note 12: The LTC1966 is a switched capacitor device and the input/output impedance is an average impedance over many clock cycles. The input impedance will not necessarily lead to an attenuation of the input signal measured. Refer to the Applications Information section titled Input Impedance for more information.

Note 13: The common mode rejection ratios of the LTC1966 are measured with DC inputs from 50mV to 350mV. The input CMRR is defined as the change in V_{IOS} measured between input levels of V_{SS} to $V_{SS} + 350mV$ and input levels of $V_{DD} - 350mV$ to V_{DD} divided by $V_{DD} - V_{SS} - 350mV$. The output CMRR is defined as the change in V_{OOS} measured with OUT RTN = V_{SS} and OUT RTN = $V_{DD} - 350mV$ divided by $V_{DD} - V_{SS} - 350mV$.

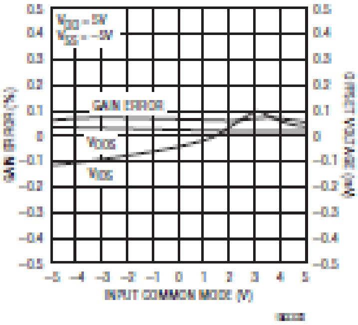
Note 14: Each input of the LTC1966 can withstand any voltage within the supply range. These inputs are protected with ESD diodes, so going beyond the supply voltages can damage the part if the absolute maximum current ratings are exceeded. Likewise for the output pins. The LTC1966 input and output voltage swings are limited by internal clipping. The maximum differential input of the LTC1966 (referred to as maximum input swing) is 1V. This applies to either input polarity, so it can be thought of as ±1V. Because the differential input voltage gets processed by the LTC1966 with gain, it is subject to internal clipping. Exceeding the 1V maximum can, depending on the input crest factor, impact the accuracy of the output voltage, but does not damage the part. Fortunately, the LTC1966's ΔΣ topology is relatively tolerant of momentary internal clipping. The input clipping is tested with a crest factor of 2, while the output clipping is tested with a DC input.

Note 15: The LTC1966 exploits oversampling and noise shaping to reduce the quantization noise of internal 1-bit analog-to-digital conversions. At higher input frequencies, increasingly large portions of this noise are aliased down to DC. Because the noise is shifted in frequency, it becomes a low frequency rumble and is only filtered at the expense of increasingly long settling times. The LTC1966 is inherently wideband, but the output accuracy is degraded by this aliased noise. These specifications apply with $C_{AVE} = 10µF$ and constitute a 3-sigma variation of the output rumble.

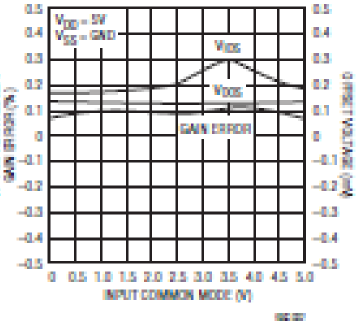
Note 16: The LTC1966 can operate down to 2.7V single supply but cannot operate at ±2.7V. This additional constraint on V_{SS} can be expressed mathematically as $-3 \bullet (V_{DD} - 2.7V) \le V_{SS} \le \text{Ground}$.

TYPICAL PERFORMANCE CHARACTERISTICS

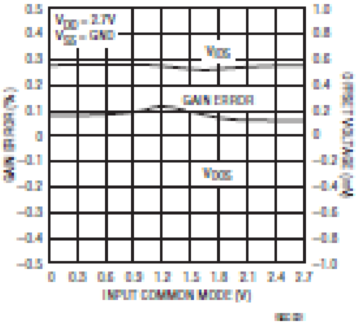
Gain and Offsets vs Input Common Mode



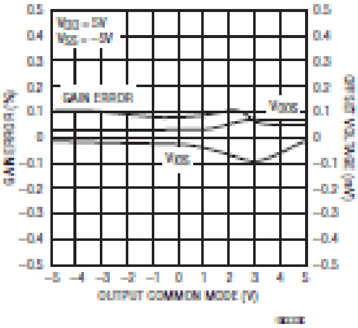
Gain and Offsets vs Input Common Mode



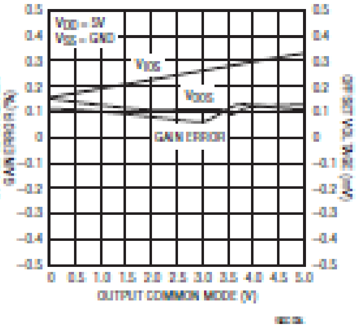
Gain and Offsets vs Input Common Mode



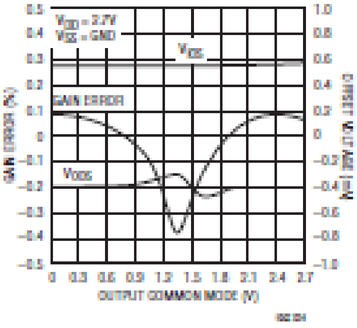
Gain and Offsets vs Output Common Mode



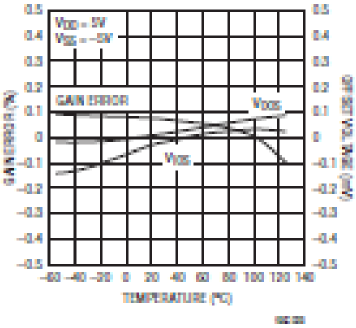
Gain and Offsets vs Output Common Mode



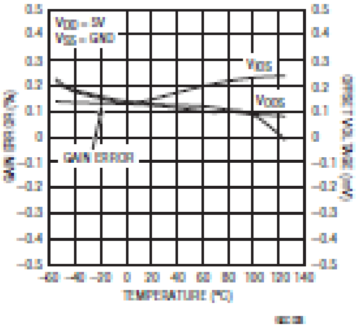
Gain and Offsets vs Output Common Mode



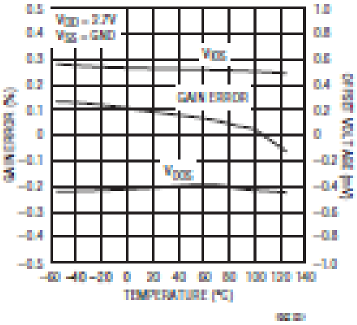
Gain and Offsets vs Temperature



Gain and Offsets vs Temperature

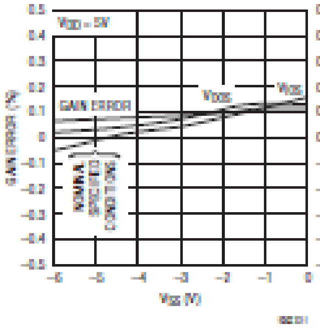


Gain and Offsets vs Temperature

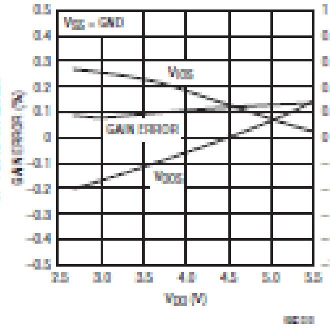


TYPICAL PERFORMANCE CHARACTERISTICS

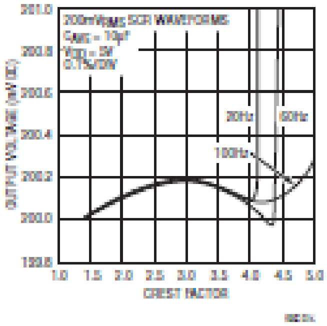
Gain and Offsets vs V_{SS} Supply



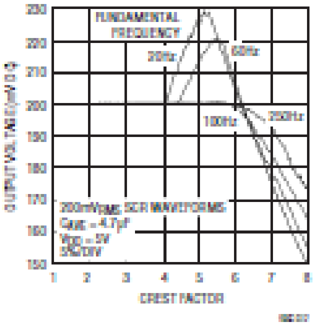
Gain and Offsets vs V_{DD} Supply



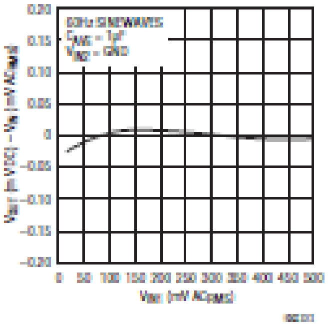
Performance vs Crest Factor



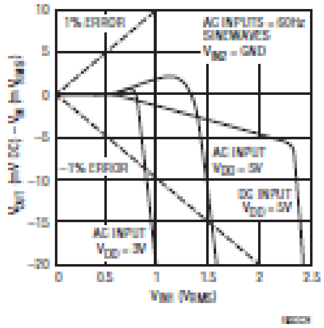
Performance vs Large Crest Factors



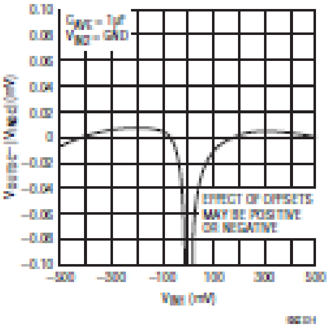
AC Linearity



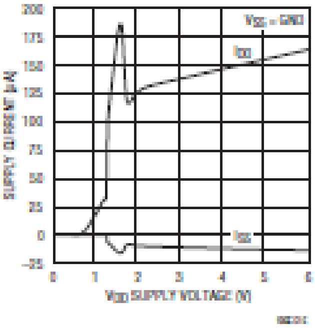
Output Accuracy vs Signal Amplitude



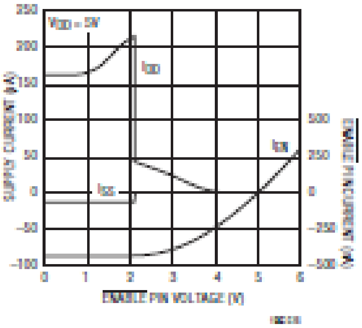
DC Linearity



Quiescent Supply Currents vs Supply Voltage

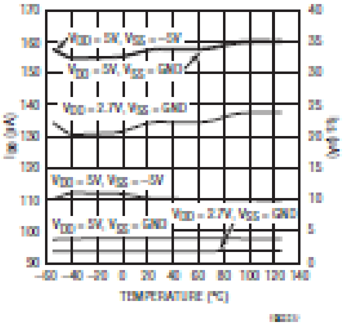


Shutdown Currents vs ENABLE Voltage

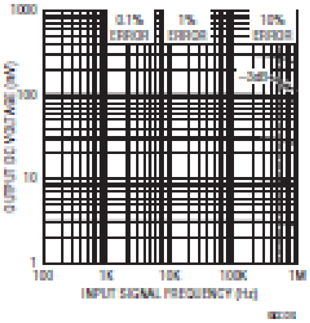


TYPICAL PERFORMANCE CHARACTERISTICS

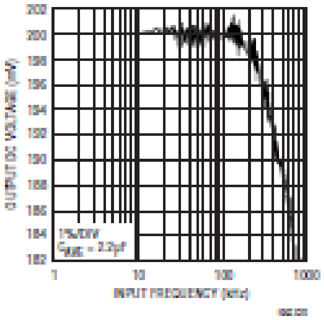
Quiescent Supply Currents vs Temperature



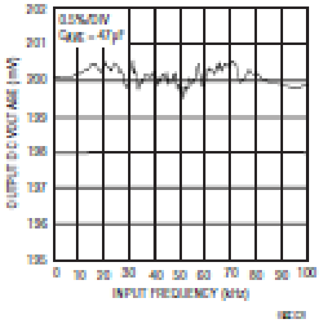
Input Signal Bandwidth



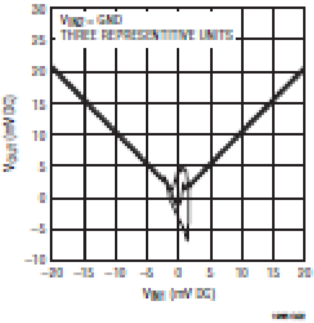
Input Signal Bandwidth



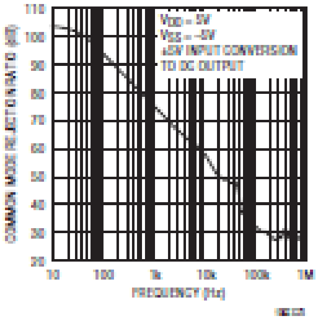
Bandwidth to 100kHz



DC Transfer Function Near Zero



Common Mode Rejection Ratio vs Frequency



PIN FUNCTIONS

GND (Pin 1): Ground. A power return pin.

IN1 (Pin 2): Differential Input. DC coupled (polarity is irrelevant).

IN2 (Pin 3): Differential Input. DC coupled (polarity is irrelevant).

V_{SS} (Pin 4): Negative Voltage Supply. GND to –5.5V.

V_{OUT} (Pin 5): Output Voltage. This is high impedance. The RMS averaging is accomplished with a single shunt capacitor from this node to OUT RTN. The transfer function is given by:

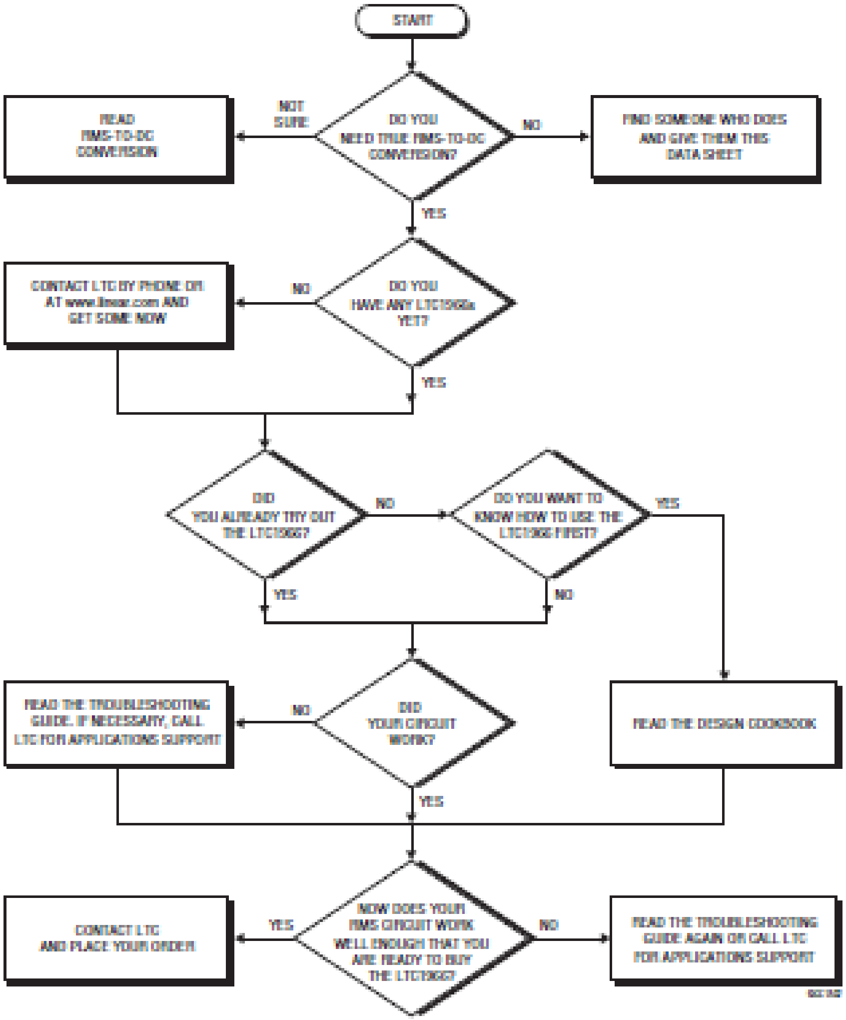
$$\left(V_{OUT} - \text{OUT RTN}\right) = \sqrt{\text{Average}\left[\left(\text{IN2} - \text{IN1}\right)^2\right]}$$

OUT RTN (Pin 6): Output Return. The output voltage is created relative to this pin. The V_{OUT} and OUT RTN pins are not balanced and this pin should be tied to a low impedance, both AC and DC. Although it is typically tied to GND, it can be tied to any arbitrary voltage, V_{SS} < OUT RTN < (V_{DD} – Max Output). Best results are obtained when OUT RTN = GND.

V_{DD} (Pin 7): Positive Voltage Supply. 2.7V to 5.5V.

$\overline{\text{ENABLE}}$ (Pin 8): An Active Low Enable Input. LTC1966 is debiased if open circuited or driven to V_{DD}. For normal operation, pull to GND, a logic low or even V_{SS}.

APPLICATIONS INFORMATION



APPLICATIONS INFORMATION

RMS-TO-DC CONVERSION

Definition of RMS

RMS amplitude is the consistent, fair and standard way to measure and compare dynamic signals of all shapes and sizes. Simply stated, the RMS amplitude is the heating potential of a dynamic waveform. A 1V_{RMS} AC waveform will generate the same heat in a resistive load as will 1V DC.

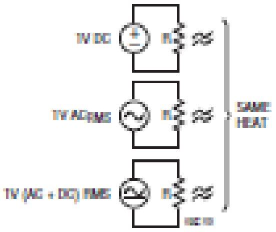


Figure 1

Mathematically, RMS is the root of the mean of the square:

$$V_{RMS} = \sqrt{\overline{V^2}}$$

Alternatives to RMS

Other ways to quantify dynamic waveforms include peak detection and average rectification. In both cases, an average (DC) value results, but the value is only accurate at the one chosen waveform type for which it is calibrated, typically sine waves. The errors with average rectification are shown in Table 1. Peak detection is worse in all cases and is rarely used.

Table 1. Errors with Average Rectification vs True RMS

WAVEFORM	V_{RMS}	AVERAGE RECTIFIED (V)	ERROR*
Square Wave	1.000	1.000	11%
Sine Wave	1.000	0.900	*Calibrate for 0% Error
Triangle Wave	1.000	0.866	–3.8%
SCR at 1/2 Power, Θ = 90°	1.000	0.637	–29.3%
SCR at 1/4 Power, Θ = 114°	1.000	0.536	–40.4%

The last two entries of Table 1 are chopped sine waves as is commonly created with thyristors such as SCRs and Triacs. Figure 2a shows a typical circuit and Figure 2b shows the resulting load voltage, switch voltage and load currents. The power delivered to the load depends on the firing angle, as well as any parasitic losses such as switch ON voltage drop. Real circuit waveforms will also typically have significant ringing at the switching transition, dependent on exact circuit parasitics. For the purposes of this data sheet, SCR waveforms refers to the ideal chopped sine wave, though the LTC1966 will do faithful RMS-to-DC conversion with real SCR waveforms as well.

The case shown is for Θ = 90°, which corresponds to 50% of available power being delivered to the load. As noted in Table 1, when Θ = 114°, only 25% of the available power is being delivered to the load and the power drops quickly as Θ approaches 180°.

With an average rectification scheme and the typical calibration to compensate for errors with sine waves, the RMS level of an input sine wave is properly reported; it is only with a nonsinusoidal waveform that errors occur. Because of this calibration, and the output reading in V_{RMS}, the term true RMS got coined to denote the use of an actual RMS-to-DC converter as opposed to a calibrated average rectifier.

Figure 2a

Figure 2b

APPLICATIONS INFORMATION

How an RMS-to-DC Converter Works

Monolithic RMS-to-DC converters use an implicit computation to calculate the RMS value of an input signal. The fundamental building block is an analog multiply/divide used as shown in Figure 3. Analysis of this topology is easy and starts by identifying the inputs and the output of the lowpass filter. The input to the LPF is the calculation from the multiplier/divider; $(V_{IN})^2/V_{OUT}$. The lowpass filter will take the average of this to create the output, mathematically:

$$V_{OUT} = \overline{\left(\frac{(V_{IN})^2}{V_{OUT}}\right)},$$

Because V_{OUT} is DC,

$$\overline{\left(\frac{(V_{IN})^2}{V_{OUT}}\right)} = \frac{\overline{\left((V_{IN})^2\right)}}{V_{OUT}}, \text{ so}$$

$$V_{OUT} = \frac{\overline{\left((V_{IN})^2\right)}}{V_{OUT}}, \text{ and}$$

$$\left(V_{OUT}\right)^2 = \overline{\left(V_{IN}\right)^2}, \text{ or}$$

$$V_{OUT} = \sqrt{\overline{\left(V_{IN}\right)^2}} = \mathrm{RMS}\left(V_{IN}\right)$$

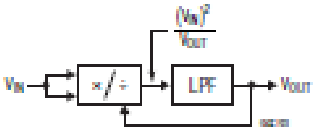


Figure 3. RMS-to-DC Converter with Implicit Computation

Unlike the prior generation RMS-to-DC converters, the LTC1966 computation does NOT use log/antilog circuits, which have all the same problems, and more, of log/antilog multipliers/dividers, i.e., linearity is poor, the bandwidth changes with the signal amplitude and the gain drifts with temperature.

How the LTC1966 RMS-to-DC Converter Works

The LTC1966 uses a completely new topology for RMS-to-DC conversion, in which a ΔΣ modulator acts as the divider, and a simple polarity switch is used as the multiplier as shown in Figure 4.

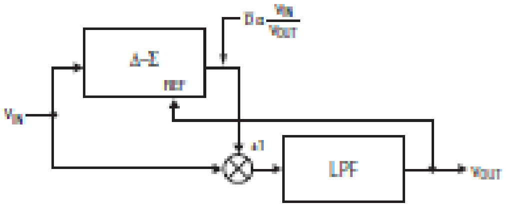


Figure 4. Topology of LTC1966

The ΔΣ modulator has a single-bit output whose average duty cycle ($\overline{D}$) will be proportional to the ratio of the input signal divided by the output. The ΔΣ is a 2nd order modulator with excellent linearity. The single bit output is used to selectively buffer or invert the input signal. Again, this is a circuit with excellent linearity, because it operates at only two points: ±1 gain; the average effective multiplication over time will be on the straight line between these two points. The combination of these two elements again creates a lowpass filter input signal proportional to $(V_{IN})^2/V_{OUT}$, which, as shown above, results in RMS-to-DC conversion.

The lowpass filter performs the averaging of the RMS function and must be a lower corner frequency than the lowest frequency of interest. For line frequency measurements, this filter is simply too large to implement on-chip, but the LTC1966 needs only one capacitor on the output to implement the lowpass filter. The user can select this capacitor depending on frequency range and settling time requirements, as will be covered in the Design Cookbook section to follow.

This topology is inherently more stable and linear than log/antilog implementations primarily because all of the signal processing occurs in circuits with high gain op amps operating closed loop.

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More detail of the LTC1966 inner workings is shown in the Simplified Schematic towards the end of this data sheet. Note that the internal scalings are such that the ΔΣ output duty cycle is limited to 0% or 100% only when V_{IN} exceeds ± 4 • V_{OUT}.

Linearity of an RMS-to-DC Converter

Linearity may seem like an odd property for a device that implements a function that includes two very nonlinear processes: squaring and square rooting.

However, an RMS-to-DC converter has a transfer function, RMS volts in to DC volts out, that should ideally have a 1:1 transfer function. To the extent that the input to output transfer function does not lie on a straight line, the part is nonlinear.

A more complete look at linearity uses the simple model shown in Figure 5. Here an ideal RMS core is corrupted by both input circuitry and output circuitry that have imperfect transfer functions. As noted, input offset is introduced in the input circuitry, while output offset is introduced in the output circuitry.

Any nonlinearity that occurs in the output circuitry will corrupt the RMS in to DC out transfer function. A nonlinearity in the input circuitry will typically corrupt that transfer function far less, simply because with an AC input, the RMS-to-DC conversion will average the nonlinearity from a whole range of input values together.

But the input nonlinearity will still cause problems in an RMS-to-DC converter because it will corrupt the accuracy as the input signal shape changes. Although an RMS-to-DC converter will convert any input waveform to a DC output, the accuracy is not necessarily as good for all waveforms as it is with sine waves. A common way to describe dynamic signal wave shapes is crest factor. The crest factor is the ratio of the peak value relative to the RMS value of a waveform. A signal with a crest factor of 4, for instance, has a peak that is four times its RMS value. Because this peak has energy (proportional to voltage squared) that is 16 times (4^2) the energy of the RMS value, the peak is necessarily present for at most 6.25% (1/16) of the time.

The LTC1966 performs very well with crest factors of 4 or less and will respond with reduced accuracy to signals with higher crest factors. The high performance with crest factors less than 4 is directly attributable to the high linearity throughout the LTC1966.

The LTC1966 does not require an input rectifier, as is common with traditional log/antilog RMS-to-DC converters. Thus, the LTC1966 has none of the nonlinearities that are introduced by rectification.

The excellent linearity of the LTC1966 allows calibration to be highly effective at reducing system errors. See System Calibration section following the Design Cookbook.

Figure 5. Linearity Model of an RMS-to-DC Converter

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DESIGN COOKBOOK

The LTC1966 RMS-to-DC converter makes it easy to implement a rather quirky function. For many applications all that will be needed is a single capacitor for averaging, appropriate selection of the I/O connections and power supply bypassing. Of course, the LTC1966 also requires power. A wide variety of power supply configurations are shown in the Typical Applications section towards the end of this data sheet.

Capacitor Value Selection

The RMS or root-mean-squared value of a signal, *the root of the mean of the square*, cannot be computed without some averaging to obtain the *mean* function. The LTC1966 true RMS-to-DC converter utilizes a single capacitor on the output to do the low frequency averaging required for RMS-to-DC conversion. To give an accurate measure of a dynamic waveform, the averaging must take place over a sufficiently long interval to average, rather than track, the lowest frequency signals of interest. For a single averaging capacitor, the accuracy at low frequencies is depicted in Figure 6.

Figure 6 depicts the so-called DC error that results at a given combination of input frequency and filter capacitor values[1]. It is appropriate for most applications, in which the output is fed to a circuit with an inherently band limited frequency response, such as a dual slope/integrating A/D converter, a ΔΣ A/D converter or even a mechanical analog meter.

However, if the output is examined on an oscilloscope with a very low frequency input, the incomplete averaging will be seen, and this ripple will be larger than the error depicted in Figure 6. Such an output is depicted in Figure 7. The ripple is at twice the frequency of the input because of the computation of the square of the input. The typical values shown, 5% peak ripple with 0.05% DC error, occur with $C_{AVE} = 1\mu F$ and $f_{INPUT} = 10Hz$.

If the application calls for the output of the LTC1966 to feed a sampling or Nyquist A/D converter (or other circuitry that will not average out this double frequency ripple) a larger averaging capacitor can be used. This trade-off is depicted in Figure 8. The peak ripple error can also be reduced by additional lowpass filtering after the LTC1966, but the simplest solution is to use a larger averaging capacitor.

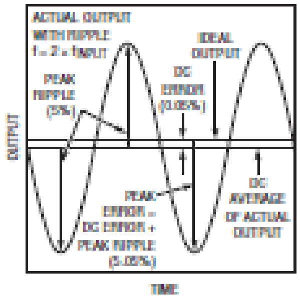

Figure 7. Output Ripple Exceeds DC Error

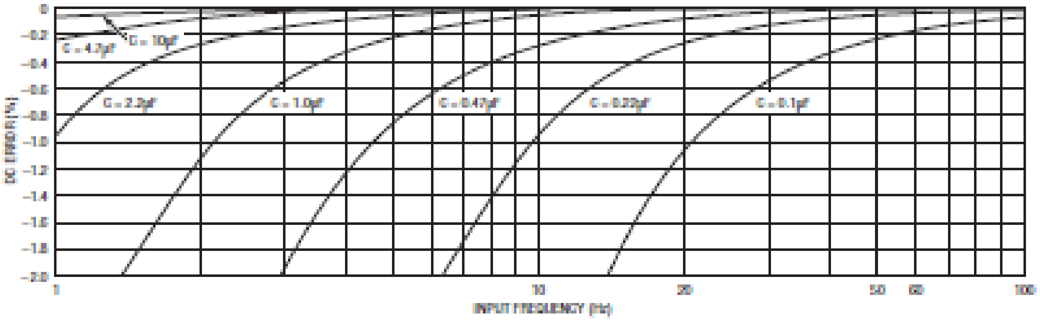

Figure 6. DC Error vs Input Frequency

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[1] This frequency dependent error is in addition to the static errors that affect all readings and are therefore easy to trim or calibrate out. The Error Analysis section to follow discusses the effect of static error terms.

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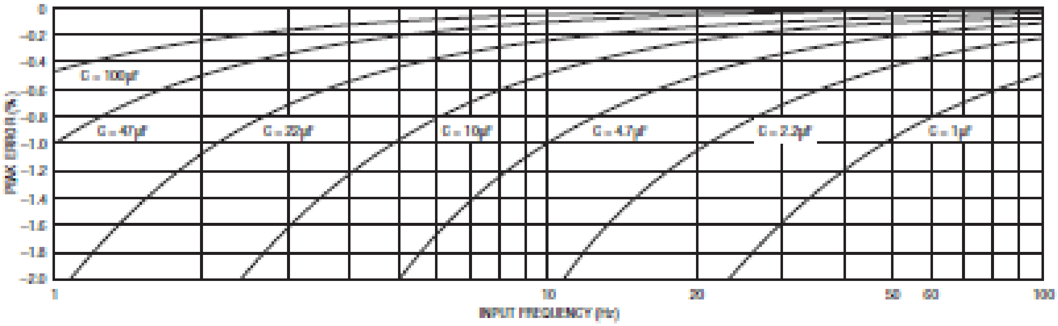

Figure 8. Peak Error vs Input Frequency with One Cap Averaging

A 1μF capacitor is a good choice for many applications. The peak error at 50Hz/60Hz will be <1% and the DC error will be <0.1% with frequencies of 10Hz or more.

Note that both Figure 6 and Figure 8 assume AC-coupled waveforms with a crest factor less than 2, such as sine waves or triangle waves. For higher crest factors and/or AC + DC waveforms, a larger C_{AVE} will generally be required. See Crest Factor and AC + DC Waveforms.

Capacitor Type Selection

The LTC1966 can operate with many types of capacitors. The various types offer a wide array of sizes, tolerances, parasitics, package styles and costs.

Ceramic chip capacitors offer low cost and small size, but are not recommended for critical applications. The value stability over voltage and temperature is poor with many types of ceramic dielectrics. This will not cause an RMS-to-DC accuracy problem except at low frequencies, where it can aggravate the effects discussed in the previous section. If a ceramic capacitor is used, it may be necessary to use a much higher nominal value in order to assure the low frequency accuracy desired.

Another parasitic of ceramic capacitors is leakage, which is again dependent on voltage and particularly temperature. If the leakage is a constant current leak, the I • R drop of the leak multiplied by the output impedance of the LTC1966 will create a constant offset of the output voltage. If the leak is Ohmic, the resistor divider formed with the LTC1966 output impedance will cause a gain error. For <0.1% gain accuracy degradation, the parallel impedance of the capacitor leakage will need to be >1000 times the LTC1966 output impedance. Accuracy at this level can be hard to achieve with a ceramic capacitor, particularly with a large value of capacitance and at high temperature.

For critical applications, a film capacitor, such as metalized polyester, will be a much better choice. Although more expensive, and larger for a given value, the value stability and low leakage make metal film capacitors a trouble free choice.

With any type of capacitor, the self resonance of the capacitor can be an issue with the switched capacitor LTC1966. If the self resonant frequency of the averaging capacitor is 1MHz or less, a second smaller capacitor should be added in parallel to reduce the impedance seen by the LTC1966 output stage at high frequencies. A capacitor 100 times smaller than the averaging capacitor will typically be small enough to be a low cost ceramic with a high quality dielectric such as X7R or NPO/COG.

Input Connections

The LTC1966 input is differential and DC coupled. The LTC1966 responds to the RMS value of the differential voltage between Pin 2 and Pin 3, including the DC portion of that difference. However, there is no DC-coupled path from the inputs to ground. Therefore, at least one of the two inputs must be connected with a DC return path to ground.

Both inputs must be connected to something. If either input is left floating, a zero volt output will result.

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For single-ended DC-coupled applications, simply connect one of the two inputs (they are interchangeable) to the signal, and the other to ground. This will work well for dual supply configurations, but for single supply configurations it will only work well for unipolar input signals. The LTC1966 input voltage range is from rail-to-rail, and when the input is driven above V_{DD} or below V_{SS} (ground for single supply operation) the gain and offset errors will increase substantially after just a few hundred millivolts of overdrive. Fortunately, most single supply circuits measuring a DC-coupled RMS value will include some reference voltage other than ground, and the second LTC1966 input can be connected to that point.

For single-ended AC-coupled applications, Figure 9 shows three alternate topologies. The first one, shown in Figure 9a uses a coupling capacitor to one input while the other is grounded. This will remove the DC voltage difference from the input to the LTC1966, and it will therefore not be part of the resulting output voltage. Again, this connection will work well with dual supply configurations, but in single supply configurations it will be necessary to raise the voltage on the grounded input to assure that the signal at the active input stays within the range of V_{SS} to V_{DD}. If there is already a suitable voltage reference available, connect the second input to that point. If not, a midsupply voltage can be created with two resistors as shown in Figure 9b.

Finally, if the input voltage is known to be between V_{SS} and V_{DD}, it can be AC-coupled by using the configuration shown in Figure 9c. Whereas the DC return path was provided through Pin 3 in Figures 9a and 9b, in this case, the return path is provided on Pin 2, through the input signal voltages. The switched capacitor action between the two input pins of the LTC1966 will cause the voltage on the coupling capacitor connected to the second input to follow the DC average of the input voltage.

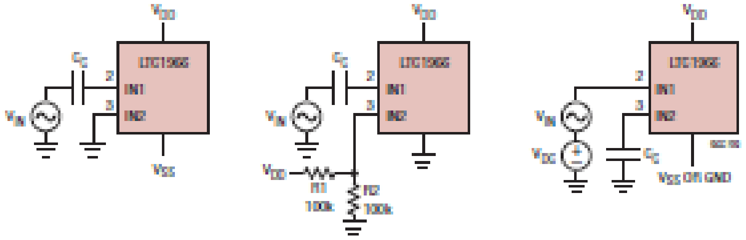
For differential input applications, connect the two inputs to the differential signal. If AC coupling is desired, one of the two inputs can be connected through a series capacitor.

In all of these connections, to choose the input coupling capacitor, C_C, calculate the low frequency coupling time constant desired, and divide by the LTC1966 differential input impedance. Because the LTC1966 input impedance is about 100 times its output impedance, this capacitor is typically much smaller than the output averaging capacitor. Its requirements are also much less stringent, and a ceramic chip capacitor will usually suffice.

Output Connections

The LTC1966 output is differentially, but not symmetrically, generated. That is to say, the RMS value that the LTC1966 computes will be generated on the output (Pin 5) relative to the output return (Pin 6), but these two pins are not interchangeable. For most applications, Pin 6 will be tied to ground (Pin 1), and this will result in the best accuracy. However, Pin 6 can be tied to any voltage between V_{SS} (Pin 4) and V_{DD} (Pin 7) less the maximum output voltage swing desired. This last restriction keeps V_{OUT} itself (Pin 5) within the range of V_{SS} to V_{DD}. If a reference level other than ground is used, it should be a low impedance, both AC and DC, for proper operation of the LTC1966.

Use of a voltage in the range of $V_{DD} – 1V$ to $V_{DD} – 1.3V$ can lead to errors due to the switch dynamics as the NMOS transistor is cut off. For this reason, it is recommended that OUT RTN = 0V if V_{DD} is ≤3V.



(9a) (9b) (9c)

Figure 9. Single-Ended AC-Coupled Input Connection Alternatives

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In any configuration, the averaging capacitor should be connected between Pins 5 and 6. The LTC1966 RMS DC output will be a positive voltage created at V_{OUT} (Pin 5) with respect to OUT RTN (Pin 6).

Power Supply Bypassing

The LTC1966 is a switched capacitor device, and large transient power supply currents will be drawn as the switching occurs. For reliable operation, standard power supply bypassing must be included. For single supply operation, a 0.01µF capacitor from V_{DD} (Pin 7) to GND (Pin 1) located close to the device will suffice. For dual supplies, add a second 0.01µF capacitor from V_{SS} (Pin 4) to GND (Pin 1), located close to the device. If there is a good quality ground plane available, the capacitors can go directly to that instead. Power supply bypass capacitors can, of course, be inexpensive ceramic types.

The sampling clock of the LTC1966 operates at approximately 200kHz, and most operations repeat at a rate of 100kHz. If this internal clock becomes synchronized to a multiple or submultiple of the input frequency, significant conversion error could occur. This is particularly important when frequencies exceeding 10kHz can be injected into the LTC1966 via supply or ground bounce. To minimize this possibility, capacitive bypassing is recommended on both supplies with capacitors placed immediately adjacent to the LTC1966. For best results, the bypass capacitors should be separately routed from Pin 7 to Pin 1, and from Pin 4 to Pin 1.

The LTC1966 needs at least 2.7V for its power supply, more for dual supply configurations. The range of allowable negative supply voltages (V_{SS}) vs positive supply voltages (V_{DD}) is shown in Figure 10. Mathematically, the V_{SS} constraint is:

$$-3 \bullet (V_{DD} - 2.7V) \leq V_{SS} \leq GND$$

The LTC1966 has internal ESD absorption devices, which are referenced to the V_{DD} and V_{SS} supplies. For effective in-circuit ESD immunity, the V_{DD} and V_{SS} pins must be connected to a low external impedance. This can be accomplished with low impedance power planes or simply with the recommended 0.01µF decoupling to ground on each supply.

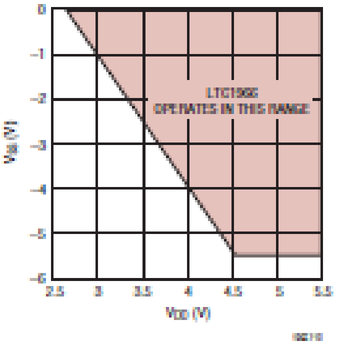

Figure 10. V_{SS} Limits vs V_{DD}

Up and Running!

If you have followed along this far, you should have the LTC1966 up and running by now! Don't forget to enable the device by grounding Pin 8, or driving it with a logic low.

Keep in mind that the LTC1966 output impedance is fairly high, and that even the standard 10MΩ input impedance of a digital multimeter (DMM) or a 10× scope probe will load down the output enough to degrade its typical gain error of 0.1%. In the end application circuit, either a buffer or another component with an extremely high input impedance (such as a dual slope integrating ADC) should be used. For laboratory evaluation, it may suffice to use a bench top DMM with the ability to disconnect the 10MΩ shunt.

If you are still having trouble, it may be helpful to skip ahead a few pages and review the Troubleshooting Guide.

What About Response Time?

With a large value averaging capacitor, the LTC1966 can easily perform RMS-to-DC conversion on low frequency signals. It compares quite favorably in this regard to prior generation products because nothing about the ΔΣ circuitry is temperature sensitive. So the RMS result doesn't get distorted by signal driven thermal fluctuations like a log/antilog circuit output does.

However, using large value capacitors results in a slow response time. Figure 11 shows the rising and falling step responses with a 1µF averaging capacitor. Although they both appear at first glance to be standard exponential

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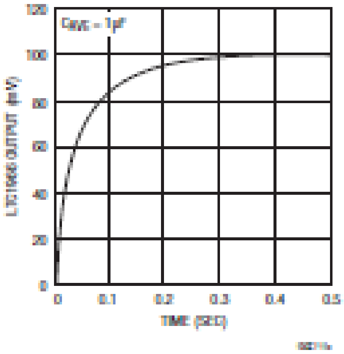

Figure 11a. LTC1966 Rising Edge with C_{AVE} = 1µF

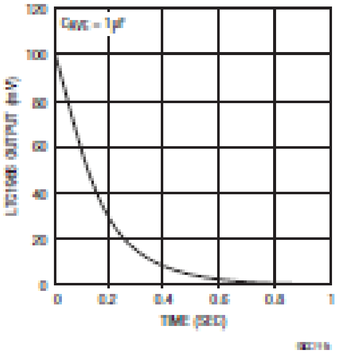

Figure 11b. LTC1966 Falling Edge with C_{AVE} = 1µF

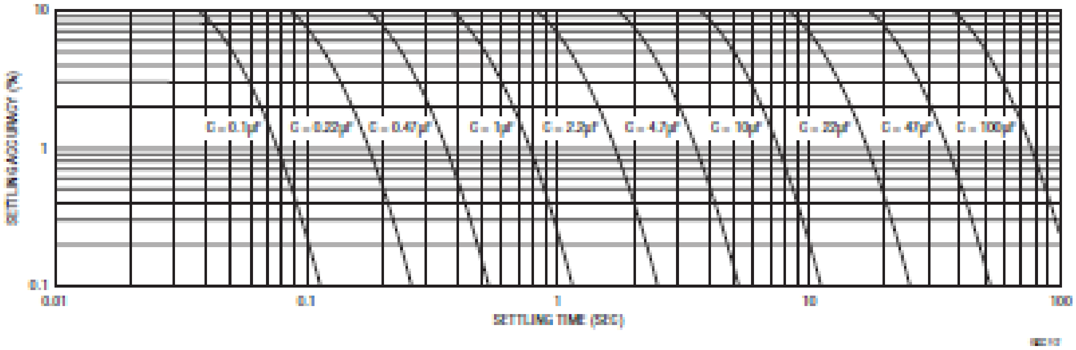

Figure 12. LTC1966 Settling Time with One Cap Averaging

decay type settling, they are not. This is due to the nonlinear nature of an RMS-to-DC calculation. Also note the change in the time scale between the two; the rising edge is more than twice as fast to settle to a given accuracy. Again this is a necessary consequence of RMS-to-DC calculation.[2]

Although shown with a step change between 0mV and 100mV, the same response shapes will occur with the LTC1966 for ANY step size. This is in marked contrast to prior generation log/antilog RMS-to-DC converters, whose averaging time constants are dependent on the signal level, resulting in excruciatingly long waits for the output to go to zero.

The shape of the rising and falling edges will be dependent on the total percent change in the step, but for less than the 100% changes shown in Figure 11, the responses will be less distorted and more like a standard exponential decay. For example, when the input amplitude is changed from 100mV to 110mV (+10%) and back (–10%), the step responses are essentially the same as a standard exponential rise and decay between those two levels. In such cases, the time constant of the decay will be in between that of the rising edge and falling edge cases of Figure 11. Therefore, the worst case is the falling edge response as it goes to zero, and it can be used as a design guide.

Figure 12 shows the settling accuracy vs settling time for a variety of averaging capacitor values. If the capacitor value previously selected (based on error requirements) gives an acceptable settling time, your design is done.

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[2] To convince oneself of this necessity, consider a pulse train of 50% duty cycle between 0mV and 100mV. At very low frequencies, the LTC1966 will essentially track the input. But as the input frequency is increased, the average result will converge to the RMS value of the input. If the rise and fall characteristics were symmetrical, the output would converge to 50mV. In fact though, the RMS value of a 100mV DC-coupled 50% duty cycle pulse train is 70.71mV, which the asymmetrical rise and fall characteristics will converge to as the input frequency is increased.

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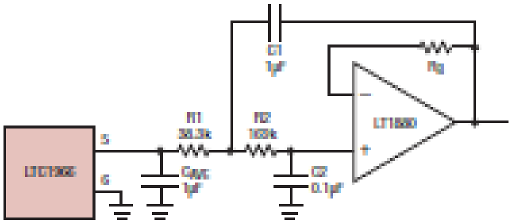
But with 100µF the settling time to even 10% is a full 38 seconds, which is a long time to wait. What can be done about such a design? If the reason for choosing 100µF is to keep the DC error with a 75mHz input less than 0.1%, the answer is: not much. The settling time to 1% of 76 seconds is just 5.7 cycles of this extremely low frequency. Averaging very low frequency signals takes a long time.

However, if the reason for choosing 100µF is to keep the peak error with a 10Hz input less than 0.05%, there is another way to achieve that result with a much improved settling time.

Reducing Ripple with a Post Filter

The output ripple is always much larger than the DC error, so filtering out the ripple can reduce the peak error substantially, without the large settling time penalty of simply increasing the averaging capacitor.

Figure 13 shows a basic 2nd order post filter, for a net 3rd order filtering of the LTC1966 RMS calculation. It uses the 85kΩ output impedance of the LTC1966 as the first resistor of a 3rd order Sallen-Key active RC filter. This topology features a buffered output, which can be desirable depending on the application. However, there are disadvantages to this topology, the first of which is that the op amp input voltage and current errors directly degrade the effective LTC1966 V_{OOS}. The table inset in Figure 13 shows these errors for four of Linear Technology's op amps.



OP AMP	LT1494	LT1880	LT1077	LT2050
LTC1966 V_{OOS}	±200µV			
V_{OS}	±375µV	±150µV	±60µV	±3µV
$I_{BOS} \cdot R$	±73µV	±323µV	±323µV	±27µV
TOTAL OFFSET	±648µV	±673µV	±583µV	±230µV
R_g VALUE	294k	SHORT	294k	SHORT
I_{SS}	1µA	1.2mA	48µA	750µA

Figure 13. Buffered Post Filter

A second disadvantage is that the op amp output has to operate over the same range as the LTC1966 output, including ground, which in single supply applications is the negative supply. Although the LTC1966 output will function fine just millivolts from the rail, most op amp output stages (and even some input stages) will not. There are at least two ways to address this. First of all, the op amp can be operated split supply if a negative supply is available. Just the op amp would need to do so; the LTC1966 can remain single supply. A second way to address this issue is to create a signal reference voltage a half volt or so above ground. This is most attractive when the circuitry that follows has a differential input, so that the tolerance of the signal reference is not a concern. To do this, tie all three ground symbols shown in Figure 13 to the signal reference, as well as to the differential return for the circuitry that follows.

Figure 14 shows an alternative 2nd order post filter, for a net 3rd order filtering of the LTC1966 RMS calculation. It also uses the 85kΩ output impedance of the LTC1966 as the first resistor of a 3rd order active RC filter, but this topology filters without buffering so that the op amp DC error characteristics do not affect the output. Although the output impedance of the LTC1966 is increased from 85kΩ to 285kΩ, this is not an issue with an extremely high input impedance load, such as a dual slope integrating ADC like the ICL7106. And it allows a generic op amp to be used, such as the SOT-23 one shown. Furthermore, it easily works on a single supply rail by tying the noninverting input of the op amp to a low noise reference as optionally shown. This reference will not change the DC voltage at the circuit output, although it does become the AC ground for the filter, thus the (relatively) low noise requirement.

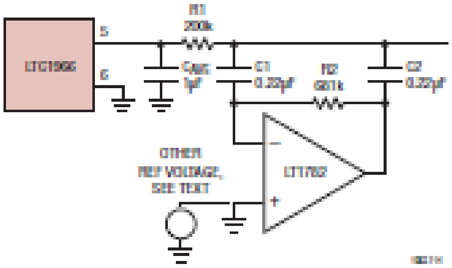


Figure 14. DC Accurate Post Filter

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Step Responses with a Post Filter

Both of the post filters, shown in Figures 13 and 14, are optimized for additional filtering with clean step responses. The 85kΩ output impedance of the LTC1966 working into a 1µF capacitor forms a 1st order LPF with a –3dB frequency of ~1.8Hz. The two filters have 1µF at the LTC1966 output for easy comparison with a 1µF only case, and both have the same relative (Bessel-like) shape. However, because of the topological differences of pole placements between the various components within the two filters, the net effective bandwidth for Figure 13 is slightly higher (≈1.2 • 1.8 ≈ 2.1Hz) than with 1µF alone, while the bandwidth for Figure 14 is somewhat lower (≈0.7 • 1.8 ≈ 1.3Hz) than with 1µF alone. To adjust the bandwidth of either of them, simply scale all the capacitors by a common multiple, and leave the resistors unchanged.

The step responses of the LTC1966 with 1µF only and with the two post filters are shown in Figure 15. This is the rising edge RMS output response to a 10Hz input starting at t = 0. Although the falling edge response is the worst case for settling, the rising edge illustrates the ripple that these post filters are designed to address, so the rising edge makes for a better intuitive comparison.

The initial rise of the LTC1966 will have enhanced slew rates with DC and very low frequency inputs due to saturation effects in the ΔΣ modulator. This is seen in Figure 15 in two ways. First, the 1µF only output is seen to rise very quickly in the first 40ms. The second way this effect shows up is that the post filter outputs have a modest overshoot, on the order of 3mV to 4mV, or 3% to 4%. This is only an issue with input frequency bursts at 50Hz or less, and even with the overshoot, the settling to a given level of accuracy improves due to the initial speedup.

As predicted by Figure 6, the DC error with 1µF is well under 1mV and is not noticeable at this scale. However, as predicted by Figure 8, the peak error with the ripple from a 10Hz input is much larger, in this case about 5mV. As can be clearly seen, the post filters reduce this ripple. Even the wider bandwidth of Figure 13's filter is seen to cut the ripple down substantially (to < 1mV) while the settling to 1% happens faster. With the narrower bandwidth of Figure 14's filter, the step response is somewhat slower, but the double frequency output ripple is just 180µV.

Figure 16 shows the step response of the same three cases with a burst of 60Hz rather than 10Hz. With 60Hz, the initial portion of the step response is free of the boost seen in Figure 15 and the two post filter responses have less than 1% overshoot. The 1µF only case still has noticeable 120Hz ripple, but both filters have removed all detectable ripple on this scale. This is to be expected; the first order filter will reduce the ripple about 6:1 for a 6:1 change in frequency, while the third order filters will reduce the ripple about 6^3:1 or 216:1 for a 6:1 change in frequency.

Again, the two filter topologies have the same relative shape, so the step response and ripple filtering trade-offs of the two are the same, with the same performance of each possible with the other by scaling it accordingly. Figures 17 and 18 show the peak error vs. frequency for a selection of capacitors for the two different filter topologies. To keep the clean step response, scale all three capacitors

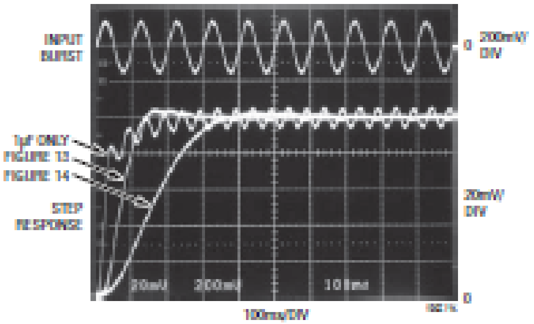

Figure 15. Step Responses with 10Hz Burst

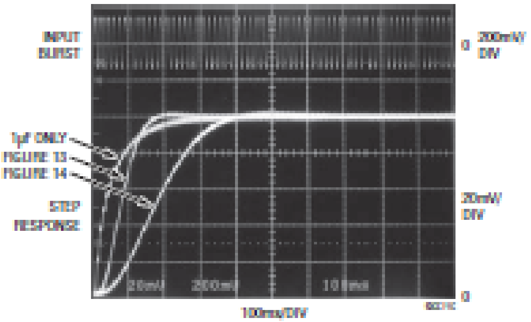

Figure 16. Step Responses with 60Hz Burst

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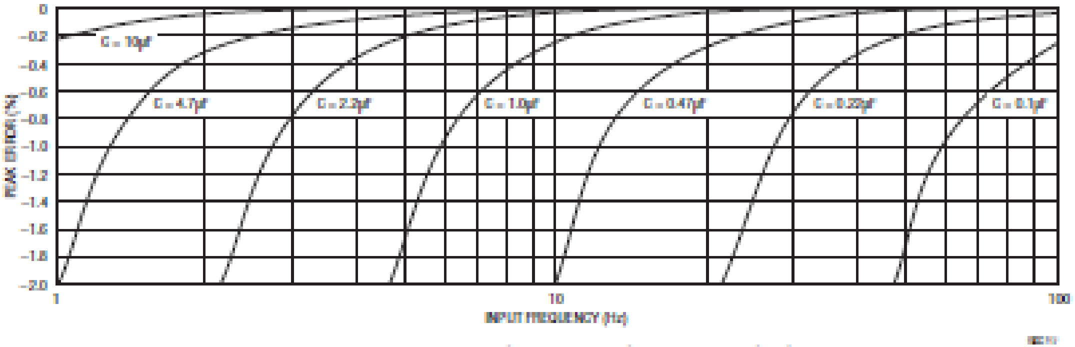

Figure 17. Peak Error vs Input Frequency with Buffered Post Filter

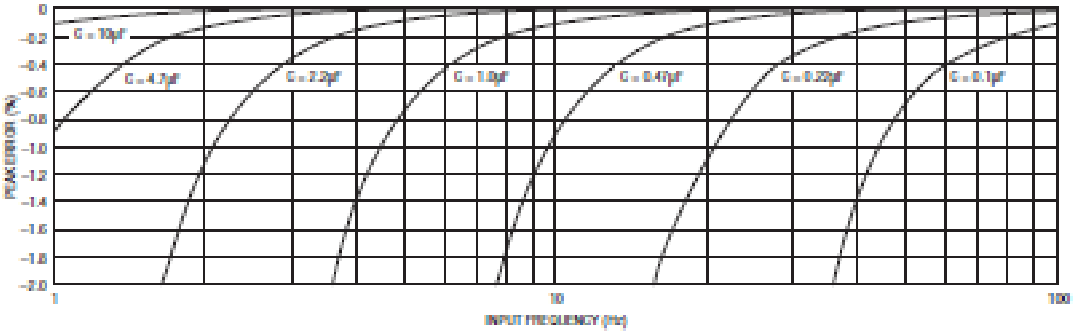

Figure 18. Peak Error vs Input Frequency with DC Accurate Post Filter

within the filter. Scaling the buffered topology of Figure 13 is simple because the capacitors are in a 10:1:10 ratio. Scaling the DC accurate topology of Figure 14 can be done with standard value capacitors; one decade of scaling is shown in Table 2.

Table 2. One Decade of Capacitor Scaling for Figure 14 with EIA Standard Values

C_{AVE}	$C_1 = C_2 =$
1µF	0.22µF
1.5µF	0.33µF
2.2µF	0.47µF
3.3µF	0.68µF
4.7µF	1µF
6.8µF	1.5µF

Figures 19 and 20 show the settling time versus settling accuracy for the buffered and DC accurate post filters, respectively. The different curves represent different scalings of the filters, as indicated by the C_{AVE} value. These are comparable to the curves in Figure 12 (single capacitor case), with somewhat less settling time for the buffered post filter, and somewhat more settling time for the DC accurate post filter. These differences are due to the change in overall bandwidth as mentioned earlier.

The other difference is the settling behavior of the filters below the 1% level. Unlike the case of a 1st order filter, any 3rd order filter can have overshoot and ringing. The filter designs presented here have minimal overshoot and ringing, but are somewhat sensitive to component mismatches. Even the ±12% tolerance of the LTC1966 output impedance can be enough to cause some ringing. The dashed lines indicate what can happen when ±5% capacitors and ±1% resistors are used.

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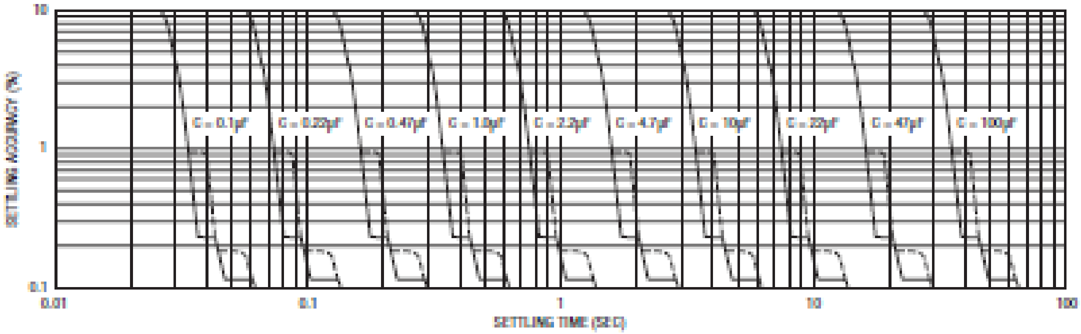

Figure 19. Settling Time with Buffered Post Filter

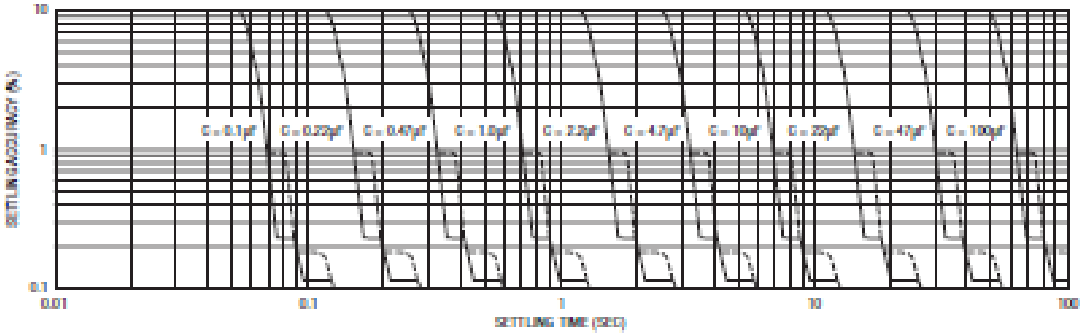

Figure 20. Settling Time with DC Accurate Post Filter

Although the settling times for the post filtered configurations shown on Figures 19 and 20 are not that much different from those with a single capacitor, the point of using a post filter is that the settling times are far better for a given level peak error. The filters dramatically reduce the low frequency averaging ripple with far less impact on settling time.

Crest Factor and AC + DC Waveforms

In the preceding discussion, the waveform was assumed to be AC-coupled, with a modest crest factor. Both assumptions ease the requirements for the averaging capacitor. With an AC-coupled sine wave, the calculation engine squares the input, so the averaging filter that follows is required to filter twice the input frequency, making its job easier. But with a sinewave that includes DC offset, the square of the input has frequency content at the input frequency and the filter must average out that lower frequency. So with AC + DC waveforms, the required value for C_{AVE} should be based on half of the lowest input frequency, using the same design curves presented in Figures 6, 8, 17 and 18.

Crest factor, which is the peak to RMS ratio of a dynamic signal, also effects the required C_{AVE} value. With a higher crest factor, more of the energy in the signal is concentrated into a smaller portion of the waveform, and the averaging has to ride out the long lull in signal activity. For busy waveforms, such as a sum of sine waves, ECG traces or SCR chopped sine waves, the required value for C_{AVE} should be based on the lowest fundamental input frequency divided as such:

$$f_{DESIGN} = \frac{f_{INPUT(MIN)}}{3 \bullet \sqrt{CF - \sqrt{2}}}$$

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using the same design curves presented in Figures 6, 8, 17 and 18. For the worst-case of square top pulse trains, that are always either zero volts or the peak voltage, base the selection on the lowest fundamental input frequency divided by twice as much:

$$f_{DESIGN} = \frac{f_{INPUT(MIN)}}{6 \bullet \sqrt{CF - \sqrt{2}}}$$

The effects of crest factor and DC offsets are cumulative. So for example, a 10% duty cycle pulse train from $0V_{PEAK}$ to $1V_{PEAK}$ ($CF = \sqrt{10} = 3.16$) repeating at 16.67ms (60Hz) input is effectively only 30Hz due to the DC asymmetry and is effectively only:

$$f_{DESIGN} = \frac{30}{6 \bullet \sqrt{3.16 - \sqrt{2}}} = 3.78\text{Hz}$$

for the purposes of Figures 6, 8, 17 and 18.

Obviously, the effect of crest factor is somewhat simplified above given the factor of 2 difference based on a subjective description of the waveform type. The results will vary somewhat based on actual crest factor and waveform dynamics and the type of filtering used. The above method is conservative for some cases and about right for others.

The LTC1966 works well with signals whose crest factor is 4 or less. At higher crest factors, the internal ΔΣ modulator will saturate, and results will vary depending on the exact frequency, shape and (to a lesser extent) amplitude of the input waveform. The output voltage could be higher or lower than the actual RMS of the input signal.

The ΔΣ modulator may also saturate when signals with crest factors less than 4 are used with insufficient averaging. This will only occur when the output droops to less than 1/4 of the input voltage peak. For instance, a DC-coupled pulse train with a crest factor of 4 has a duty cycle of 6.25% and a $1V_{PEAK}$ input is $250mV_{RMS}$. If this input is 50Hz, repeating every 20ms, and $C_{AVE} = 1\mu F$, the output will droop during the inactive 93.75% of the waveform. This droop is calculated as:

$$V_{MIN} = \frac{V_{RMS}}{2}\left(1 - e^{-\left(\frac{\text{INACTIVE TIME}}{2 \bullet Z_{OUT} \bullet C_{AVE}}\right)}\right)$$

For the LTC1966, whose output impedance (Z_{OUT}) is 85kΩ, this droop works out to –5.22%, so the output would be reduced to 237mV at the end of the inactive portion of the input. When the input signal again climbs to $1V_{PEAK}$, the peak/output ratio is 4.22.

With $C_{AVE} = 10\mu F$ the droop is only –0.548% to 248.6mV and the peak/output ratio is just 4.022, which the LTC1966 has enough margin to handle without error.

For crest factors less than 3.5, the selection of C_{AVE} as previously described should be sufficient to avoid this droop and modulator saturation effect. But with crest factors above 3.5, the droop should also be checked for each design.

Error Analyses

Once the RMS-to-DC conversion circuit is working, it is time to take a step back and do an analysis of the accuracy of that conversion. The LTC1966 specifications include three basic static error terms, V_{OOS}, V_{IOS} and GAIN. The output offset is an error that simply adds to (or subtracts from) the voltage at the output. The conversion gain of the LTC1966 is nominally 1.000 V_{DCOUT}/V_{RMSIN} and the gain error reflects the extent to which this conversion gain is not perfectly unity. Both of these affect the results in a fairly obvious way.

Input offset on the other hand, despite its conceptual simplicity, effects the output in a nonobvious way. As its name implies, it is a constant error voltage that adds directly with the input. And it is the sum of the input and V_{IOS} that is RMS converted.

This means that the effect of V_{IOS} is warped by the nonlinear RMS conversion. With 0.2mV (typ) V_{IOS}, and a $200mV_{RMS}$ AC input, the RMS calculation will add the DC and AC terms in an RMS fashion and the effect is negligible:

$$\begin{aligned} V_{OUT} &= \sqrt{(200\text{mV AC})^2 + (0.2\text{mV DC})^2} \\ &= 200.0001\text{mV} \\ &= 200\text{mV} + 1/2\text{ppm} \end{aligned}$$

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But with 10× less AC input, the error caused by V_{IOS} is 100× larger:

$$V_{OUT} = \sqrt{(20mV\ AC)^2 + (0.2mV\ DC)^2}$$
$$= 20.001mV$$
$$= 20mV + 50ppm$$

This phenomena, although small, is one source of the LTC1966's residual nonlinearity.

On the other hand, if the input is DC-coupled, the input offset voltage adds directly. With +200mV and a +0.2mV V_{IOS}, a 200.2mV output will result, an error of 0.1% or 1000ppm. With DC inputs, the error caused by V_{IOS} can be positive or negative depending if the two have the same or opposing polarity.

The total conversion error with a sine wave input using the typical values of the LTC1966 static errors is computed as follows:

$$V_{OUT} = (\sqrt{(500mV\ AC)^2 + (0.2mV\ DC)^2}) \bullet 1.001 + 0.1mV$$
$$= 500.600mV$$
$$= 500mV + 0.120\%$$

$$V_{OUT} = (\sqrt{(50mV\ AC)^2 + (0.2mV\ DC)^2}) \bullet 1.001 + 0.1mV$$
$$= 50.150mV$$
$$= 50mV + 0.301\%$$

$$V_{OUT} = (\sqrt{(5mV\ AC)^2 + (0.2mV\ DC)^2}) \bullet 1.001 + 0.1mV$$
$$= 5.109mV$$
$$= 5mV + 2.18\%$$

As can be seen, the gain term dominates with large inputs, while the offset terms become significant with smaller inputs. In fact, 5mV is the minimum RMS level needed to keep the LTC1966 calculation core functioning normally, so this represents the worst-case of usable input levels.

Using the worst-case values of the LTC1966 static errors, the total conversion error is:

$$V_{OUT} = (\sqrt{(500mV\ AC)^2 + (0.8mV\ DC)^2}) \bullet 1.003 + 0.2mV$$
$$= 501.70mV$$
$$= 500mV + 0.340\%$$

$$V_{OUT} = (\sqrt{(50mV\ AC)^2 + (0.8mV\ DC)^2}) \bullet 1.003 + 0.2mV$$
$$= 50.356mV$$
$$= 50mV + 0.713\%$$

$$V_{OUT} = (\sqrt{(5mV\ AC)^2 + (0.8mV\ DC)^2}) \bullet 1.003 + 0.2mV$$
$$= 5.279mV$$
$$= 5mV + 5.57\%$$

These static error terms are in addition to dynamic error terms that depend on the input signal. See the Design Cookbook for a discussion of the DC conversion error with low frequency AC inputs. The LTC1966 bandwidth limitations cause additional errors with high frequency inputs. Another dynamic error is due to crest factor. The LTC1966 performance versus crest factor is shown in the Typical Performance Characteristics.

Monotonicity and Linearity

The LTC1966, like all implicit RMS-to-DC convertors (Figure 3), has a division with the output in the denominator. This works fine most of the time, but when the output is zero or near zero this becomes problematic. The LTC1966 has multiple switched capacitor amplifier stages, and depending on the different offsets and their polarity, the DC transfer curve near zero input can take a few different forms, as shown in the Typical Performance Characteristics graph titled DC Transfer Function Near Zero.

Some units (about 1 of every 16) will even be well behaved with a transfer function that is the upper half of a unit rectangular hyperbola with a focal point on the y-axis of a few millivolts.[3] For AC inputs, these units will have a monotonic transfer function all the way down to zero input.

The LTC1966 is trimmed for offsets as small as practical, and the resulting behavior is the best statistical linearity provided the zero region troubles are avoided.

It is possible, and even easy, to force the zero region to be well behaved at the price of additional (though predictable) V_{OOS} and some linearity error. For large enough input signals, this linearity error may be negligible.

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[3] In general, every LTC1966 will have a DC transfer function that is essentially a unit rectangular hyperbola (the gain is not always exactly unity, but the gain error is small) with an X- and Y- offset equal to V_{IOS} and V_{OOS}, respectively, until the inputs are small enough that the delta sigma section gets confused. While some units will be the north half of a north south pair, other units will have two upper halfs of the conjugate, east west, hyperbolas. The circuit of Figure 23 will assure a continuous transfer function.

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To do this, inject current into the output. As shown in Figure 21, the charge pump output impedance is 170kΩ, with the computational feedback cutting the closed loop output impedance to the 85kΩ specification. By injecting 30nA of current into this 170Ω, with zero input, a 5mV offset is created at the output feedback point, which is sufficient to overcome the 5mV minimum signal level. With large enough input signals, the computational feedback cuts the output impedance to 85kΩ so the transfer function asymptotes will have an output offset of 2.5mV, as shown in Figure 22. This is the additional, predictable, V_{OOS} that is added, and should be subtracted from the RMS results, either digitally, or by an analog means.

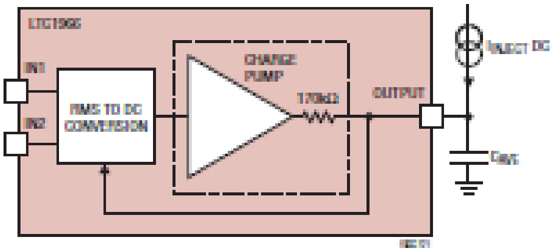

Figure 21. Behavioral Block Diagram of LTC1966

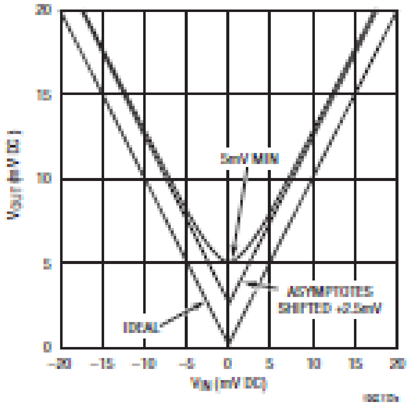

Figure 22a. DC Transfer Function with I_{INJECT} = 30nA

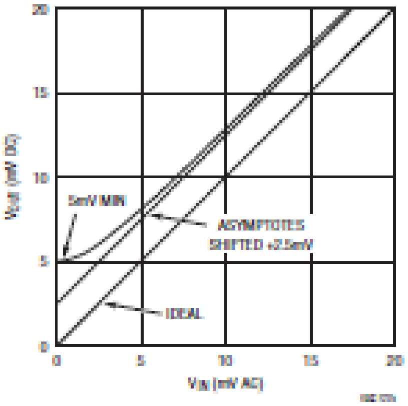

Figure 22b. AC Transfer Function with I_{INJECT} = 30nA

Figure 23 shows an analog implementation of this with the offset and gain errors corrected; only the slight, but necessary, degradation in nonlinearity remains. The circuit works by creating approximately 300mV of bias at the junction of the 10MΩ resistors when the LTC1966's input/output are zero. The 10MΩ resistor to the LTC1966 output therefore feeds in 30nA. The loading of this resistor causes a slight reduction in gain which is corrected, as is the nominal 2.5mV offset, by the LT1494 op amp.

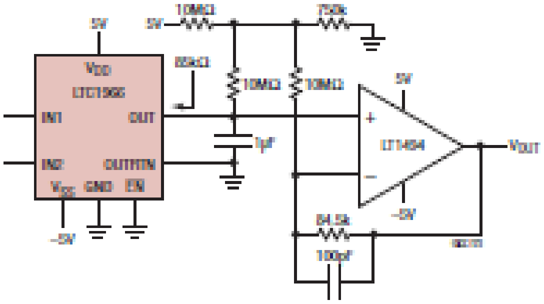

Figure 23. Monotonic AC Response with Offset and Gain Corrected

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The two 10MΩ resistors not connected to the supply can be any value as long as they match and the feed voltage is changed for 30nA injection. The op amp gain is only 1.00845, so the output is dominated by the LTC1966 RMS results, which keeps errors low. With the values shown, the resistors can be ±2% and only introduce ±170ppm of gain error. The 84.5k resistor is the closest match in the 1% EIA values but if the 2% EIA value of 82k were used instead, the gain would only be reduced by 248ppm.

This low error sensitivity is important because the LTC1966 output impedance is 85kΩ ±11.8%, which can create a gain error of ±0.1%; enough to degrade the overall gain accuracy somewhat. This gain variation term is increased with lower value feed resistors, and decreased with higher value feed resistors.

A bigger error caused by the variation of the LTC1966 output impedance is imperfect cancelation of the output offset introduced by the injected current. The offset correction provided by the LT1494 will be based on a consistent 84.5kΩ times the injected current, while the LTC1966 output impedance will vary enough that the output offset will have a ±300μV range about the nominal 2.5mV. If this level of output offset is not acceptable, either system calibration or a potentiometer in the LT1494 feedback may be needed.

If the two 10MΩ feed resistors to the LT1494 have significant mismatch, cancellation of the 2.5mV offset would be further impacted, so it is probably worth paying an extra penny or so for 1% resistors or even the better temperature stability of thin film devices. The 300mV feed voltage is not particularly critical because it is nominally cancelled, but the offset errors due to these resistance mismatches is scaled by that voltage.

Note that the input bias current of the op amp used in Figure 23 is also nominally cancelled, but it will add or subtract to the total current injected into the LTC1966 output. With the 1nA I_{BIAS} of the LT1494 this is negligible. While it is possible to eliminate the feed resistors by using an op amp with a PNP input stage whose I_{BIAS} is 30nA or more, I_{BIAS} is usually only specified for maximum and this circuit needs a minimum of 30nA, therefore such an approach may not always work.

Because the circuit of Figure 23 subtracts the offset created by the injected current, the LT1494 output with zero LTC1966 input will rest at +2.5mV, nominal before offsets, rather then the 5mV seen in Figure 22.

Output Errors Versus Frequency

As mentioned in the Design Cookbook, the LTC1966 performs very well with low frequency and very low frequency inputs, provided a large enough averaging capacitor is used.

However, the LTC1966 will have additional dynamic errors as the input frequency is increased. The LTC1966 is designed for high accuracy RMS-to-DC conversion of signals into the audible range. The input sampling amplifiers have a –3dB frequency of 800kHz or so. However, the switched capacitor circuitry samples the inputs at a modest 100kHz nominal. The response versus frequency is depicted in the Typical Performance Characteristics titled Input Signal Bandwidth. Although there is a pattern to the response versus frequency that repeats every sample frequency, the errors are not overwhelming. This is because LTC1966 RMS calculation is inherently wideband, operating properly with minimal oversampling, or even undersampling, using several proprietary techniques to exploit the fact that the RMS value of an aliased signal is the same as the RMS value of the original signal. However, a fundamental feature of the ΔΣ modulator is that sample estimation noise is shaped such that minimal noise occurs with input frequencies much less than the sampling frequency, but such noise peaks when input frequency reaches half the sampling frequency. Fortunately the LTC1966 output averaging filter greatly reduces this error, but the RMS-to-DC topology frequency shifts the noise to low (baseband) frequencies. So with input frequencies above 5kHz to 10kHz, the output will slowly wander around ±a few percent.

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Input Impedance

The LTC1966 true RMS-to-DC converter utilizes a 2.5pF capacitor to sample the input at a nominal 100kHz sample frequency. This accounts for the 8MΩ input impedance. See Figure 24 for the equivalent analog input circuit. Note however, that the 8MΩ input impedance does not directly affect the input sampling accuracy. For instance, if a 100k source resistance is used to drive the LTC1966, the sampling action of the input stage will drag down the voltage seen at the input pins with small spikes at every sample clock edge as the sample capacitor is connected to be charged. The time constant of this combination is small, 2.5pF • 100kΩ = 250ns, and during the 2.5µs period devoted to sampling, ten time constants elapse. This allows each sample to settle to within 46ppm and it is these samples that are used to compute the RMS value.

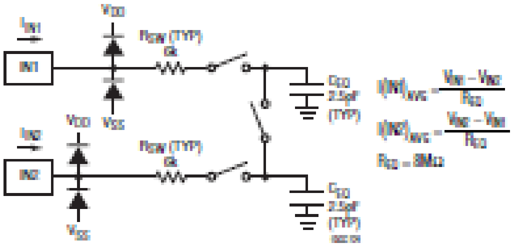

Figure 24. LTC1966 Equivalent Analog Input Circuit

This is a much higher accuracy than the LTC1966 conversion limits, and far better than the accuracy computed via the simplistic resistive divider model:

$$\begin{aligned} V_{IN} &= V_{SOURCE}\frac{R_{IN}}{R_{IN}+R_{SOURCE}} \\ &= V_{SOURCE}\frac{8M\Omega}{8M\Omega+100k\Omega} \\ &= V_{SOURCE}-1.25\% \end{aligned}$$

This resistive divider calculation does give the correct model of what voltage is seen at the input terminals by a parallel load averaged over a several clock cycles, which is what a large shunt capacitor will do—average the current spikes over several clock cycles.

When high source impedances are used, care must be taken to minimize shunt capacitance at the LTC1966 input so as not to increase the settling time. Shunt capacitance of just 2.5pF will double the input settling time constant and the error in the above example grows from 46ppm to 0.67% (6700ppm). A 13pF scope probe will increase the error to almost 20%. As a consequence, it is important to *not* try to filter the input with large input capacitances unless driven by a low impedance. Keep time constant ≪ 2.5µs.

When the LTC1966 is driven by op amp outputs, whose low DC impedance can be compromised by sharp capacitive load switching, a small series resistor may be added. A 10k resistor will easily settle with the 2.5pF input sampling capacitor to within 1ppm.

These are important points to consider both during design and debug. During lab debug, and even production testing, a high value series resistor to any test point is advisable.

Output Impedance

The LTC1966 output impedance during operation is similarly due to a switched capacitor action. In this case, 59pF of on-chip capacitance operating at 100kHz translates into 170kΩ. The closed loop RMS-to-DC calculation cuts that in half to the nominal 85kΩ specified.

In order to create a DC result, a large averaging capacitor is required. Capacitive loading and time constants are not an issue on the output.

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However, resistive loading is an issue and the 10MΩ impedance of a DMM or 10× scope probe will drag the output down by –0.85% typ.

During shutdown, the switching action is halted and a fixed 30k resistor shunts V_{OUT} to OUT RTN so that C_{AVE} is discharged.

Guard Ringing the Output

The LTC1966's combination of precision and high output impedance can present challenges that make the use of a guard ring around the output a good idea for many applications.

As mentioned above, a 10M resistive loading to ground will drag down the gain far more than the specificed gain tolerance. On a printed circuit board, contaminants from solder flux residue to finger grime can create parasitic resistances, which may be very high impedance, but can have deleterious effects on the realized accuracy. As an example, if the output (Pin 5) is routed near V_{SS} (Pin 4) in a ±5V application, a parasitic resistance of 1G (1,000M) is enough to introduce a –425µV output offset error, more than the specified limit of the LTC1966 itself.

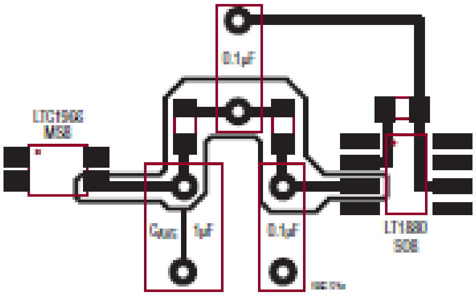


Figure 24a. PCB Layout of Figure 13 with Guard Ring

Use of a guard ring, wherein the LTC1966 output node is completely surrounded by a low impedance voltage, can reduce leakage related errors substantially. The ground ring can be tied to OUTRTN (Pin 6) and should encircle the output (Pin 5), the averaging capacitor terminal, and the destination terminal at the ADC, filter op amp, or whatever else may be next.

Figure 24a shows a sample PCB layout for the circuit of Figure 13, wherein the guard ring trace encloses R1, R2, and the terminals of C1, C2, and the op amp input connected to the high impedance LTC1966 Output. For the circuit of figure 14, the guard ring should enclose R1 and the terminals of C1 and C2, as well as the terminal at the ultimate destination.

Figure 24b shows a sample PCB layout for the circuit of Figure 23. The summing node of the LT1494 has the same high impedance and high accuracy as the LTC1966 output, so here the guard ring encircles both of them. Any leakage between them is benign because the LT1494 forces them to the same nominal voltage.

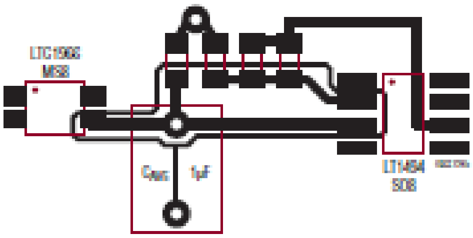


Figure 24b. PCB Layout of Figure 23 with Guard Ring

APPLICATIONS INFORMATION

Interfacing with an ADC

The LTC1966 output impedance and the RMS averaging ripple need to be considered when using an analog-to-digital converter (ADC) to digitize the LTC1966 RMS result.

The simplest configuration is to connect the LTC1966 directly to the input of a type 7106/7136 ADC as shown in Figure 25a. These devices are designed specifically for DVM/DPM use and include display drivers for a 3 1/2 digit LCD segmented display. Using a dual slope conversion, the input is sampled over a long integration window, which results in rejection of line frequency ripple when integration time is an integer number of line cycles. Finally, these parts have an input impedance in the GΩ range, with specified input leakage of 10pA to 20pA. Such a leakage, combined with the LTC1966 output impedance, results in just 1μV to 2μV of additional output offset voltage.

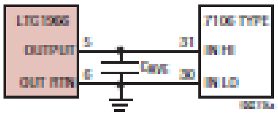


Figure 25a. Interfacing to DVM/DPM ADC

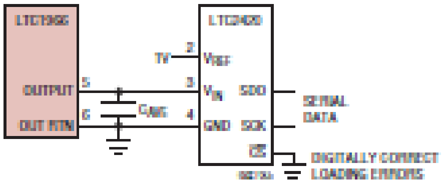


Figure 25b. Interfacing to LTC2420

Another type of ADC that has inherent rejection of RMS averaging ripple is an oversampling ΔΣ. With most, but not all, of these devices, it is possible to connect the LTC1966 output directly to the converter input. Issues to look out for are the input impedance, and any input sampling currents. The input sampling currents drawn by ΔΣ ADCs often have large spikes of current with short durations that can confuse some op amps, but with the large C_{AVE} needed by the LTC1966 these are not an issue.

The average current is important, as it can create LTC1966 errors; if it is constant it will create an offset, while average currents that change with the voltage level create gain errors. Some converters run continuously, others only sample upon demand, and this will change the results in ways that need to be understood. The LTC1966 output impedance has a loose tolerance relative to the usual resistors and the same can be true for the input impedance of ΔΣ ADC, resulting in gain errors from part-to-part. The system calibration techniques described in the following section should be used in applications that demand tight tolerances.

One example of driving an oversampling ΔΣ ADC is shown in Figure 25b. In this circuit, the LTC2420 is used with a 1V V_{REF}. Since the LTC1966 output voltage range is about 1V, and the LTC2420 has a ±12.5% extended input range, this configuration matches the two ranges with room to spare. The LTC2420 has an input impedance of 16.6MΩ, resulting in a gain error of –0.4% to –0.6%. In fact, the LTC2420 DC input current is not zero at 0V, but rather at one half its reference, so both an output offset and a gain error will result. These errors will vary from part to part, but with a specific LTC1966 and LTC2420 combination, the errors will be fixed, varying less than ±0.05% over temperature. So a system that has digital calibration can be quite accurate despite the nominal gain and offset error. With 20 bits of resolution, this part is more accurate than the LTC1966, but the extra resolution is helpful because it reduces nonlinearity at the LSB transitions as a digital gain correction is made. Furthermore, its small size and ease of use make it attractive.

APPLICATIONS INFORMATION

As is shown in Figure 25b, where the LTC2420 is set to continuously convert by grounding the CS pin. The gain error will be less if CS is driven at a slower rate, however, the rate should either be consistent or at a rate low enough that the LTC1966 and its output capacitor have fully settled by the beginning of each conversion, so that the loading errors are consistent.

Note that in this circuit, the input current of the LTC2420 is being used to assure monotonicity. The LTC2420 Z_{IN} of 16.6MΩ is effectively connected to half the reference voltage, so when the LTC1966 has zero signal, 500mV/16.6MΩ = 30nA is provided.

Alternatively, a 5V V_{REF} can be used, but in this case the LTC1966 output span will only use 20% of the LTC2420's input voltage range. Furthermore, if the OUTRTN remains grounded, the injected current with zero signal will be 150nA, resulting in 5× the offset error and nonlinearity shown in Figure 22.

In both of the circuits of Figure 25, a guard ring only has to encircle three terminals, the LTC1966 output, the top of the averaging capacitor, and the ADC input. Figure 26 shows the top copper patterns for example PCB layouts of each.

The low power consumption of the LTC1966 makes it well suited for battery powered applications, and its slow output (DC) makes it an ideal candidate for a micropower ADC.

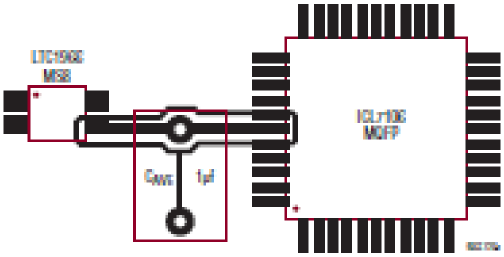


Figure 26a. PCB Layout of Figure 25a with Guard Ring

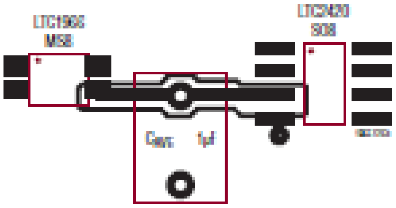


Figure 26b. PCB Layout of Figure 25b with Guard Ring

Figure 10 in Application Note 75, for instance, details a 10-bit ADC with a 35ms conversion time that uses just 29µA of supply current. Such an ADC may also be of use within a 4mA to 20mA loop.

Other types of ADCs sample the input signal once and perform a conversion on that one sample. With these ADCs (Nyquist ADCs), a post filter will be needed in most cases to reduce the peak error with low input frequencies. The DC accurate filter of Figure 14 is attractive from an error standpoint, but it increases the impedance at the ADC input. In most cases, the buffered post filter of Figure 13 will be more appropriate for use with Nyquist analog-to-digital converters.

SYSTEM CALIBRATION

The LTC1966 static accuracy can be improved with end system calibration. Traditionally, calibration has been done at the factory, or at a service depot only, typically using manually adjusted potentiometers. Increasingly, systems are being designed for electronic calibration where the accuracy corrections are implemented in digital code wherever possible, and with calibration DACs where necessary. Additionally, many systems are now designed for self calibration, in which the calibration occurs inside the machine, automatically without user intervention.

APPLICATIONS INFORMATION

Whatever calibration scheme is used, the linearity of the LTC1966 will improve the calibrated accuracy over that achievable with older log/antilog RMS-to-DC converters. Additionally, calibration using DC reference voltages are essentially as accurate with the LTC1966 as those using AC reference voltages. Older log/antilog RMS-to-DC converters required nonlinear input stages (rectifiers) whose linearity would typically render DC based calibration unworkable.

The following are four suggested calibration methods. Implementations of the suggested adjustments are dependent on the system design, but in many cases, gain and output offset can be corrected in the digital domain, and will include the effect of all gains and offsets from the LTC1966 output through the ADC. Input offset voltage, on the other hand, will have to be corrected with adjustment to the actual analog input to the LTC1966.

The methods below assume the unaltered linearity of the LTC1966, i.e. without the monotonicity fix of Figure 21. If this is present, the V_{OOS} shift it introduces should be taken out before using either method for which V_{OOS} is not calibrated. Also, the nonlinearity it introduces will increase the 20mV readings discussed below by 0.78% but increase the 200mV readings only 78ppm. There are a variety of ways to deal with these errors, including possibly ignoring them, but the specifics will depend on system requirements. Designers are cautioned to avoid the temptation to digitally take out the hyperbolic transfer function introduced because if the offsets are not exactly the nominals assumed, the system will end up right back where it began with a potential discontinuity with zero input, either from a divide by zero or from a square root of a negative number in the calculations to undo the hyperobic transfer function. An adaptive algorithm would most likely be necessary to safely take out more than half of the introduced nonlinearity.

If a 5V reference is used in the connection of Figure 25b, the V_{OOS} and nonlinearity created would be even larger, and will no doubt be more tempting to correct for. Designers are likewise cautioned against correcting for all of the nonlinearity.

AC-Only, 1 Point

The dominant error at full-scale will be caused by the gain error, and by applying a full-scale sine wave input, this error can be measured and corrected for. Unlike older log/antilog RMS-to-DC converters, the correction should be made for zero error at full scale to minimize errors throughout the dynamic range.

The best frequency for the calibration signal is roughly ten times the –0.1% DC error frequency. For 1µF, –0.1% DC error occurs at 8Hz, so 80Hz is a good calibration frequency, although anywhere from 60Hz to 100Hz should suffice.

The trade-off here is that on the one hand, the DC error is input frequency dependent, so a calibration signal frequency high enough to make the DC error negligible should be used. On the other hand, as low a frequency as can be used is best to avoid attenuation of the calibrated AC signal, either from parasitic RC loading or insufficient op amp gain. For instance, with a 1kHz calibration signal, a 1MHz op amp will typically only have 60dB of open loop gain, so it could attenuate the calibration signal a full 0.1%.

AC-Only, 2 Point

The next most significant error for AC-coupled applications will be the effect of output offset voltage, noticeable at the bottom end of the input scale. This too can be calibrated out if two measurements are made, one with a full-scale sine wave input and a second with a sine wave input (of the same frequency) at 10% of full-scale. The trade-off in selecting this second level is that it should be small enough that the gain error effect becomes small compared to the gain error effect at full-scale, while on the other hand, not using so small an input that the input offset voltage becomes an issue.

APPLICATIONS INFORMATION

The calculations of the error terms for a 200mV full-scale case are:

$$\text{Gain} = \frac{\text{Reading at 200mV} - \text{Reading at 20mV}}{\text{180mV}}$$

$$\text{Output Offset} = \frac{\text{Reading at 20mV}}{\text{Gain}} - \text{20mV}$$

DC, 2 Point

DC based calibration is preferable in many cases because a DC voltage of known, good accuracy is easier to generate than such an AC calibration voltage. The only down side is that the LTC1966 input offset voltage plays a role. It is therefore suggested that a DC based calibration scheme check at least two points: ±full-scale. Applying the –full-scale input can be done by physically inverting the voltage or by applying the same +full-scale input to the opposite LTC1966 input.

For an otherwise AC-coupled application, only the gain term may be worth correcting for, but for DC-coupled applications, the input offset voltage can also be calculated and corrected for.

The calculations of the error terms for a 200mV full-scale case are:

$$\text{Gain} = \frac{\text{Reading at 200mV} + \text{Reading at} - \text{200mV}}{\text{400mV}}$$

$$\text{Input Offset} = \frac{\text{Reading at} - \text{200mV} - \text{Reading at 200mV}}{2 \bullet \text{Gain}}$$

Note: Calculation of and correction for input offset voltage are the only way in which the two LTC1966 inputs (IN1, IN2) are distinguishable from each other. The calculation above assumes the standard definition of offset; that a positive offset is the case of a positive voltage error inside the device that must be corrected by applying a like negative voltage outside. The offset is referred to whichever pin is driven positive for the +full-scale reading.

DC, 3 Point

One more point is needed with a DC calibration scheme to determine output offset voltage: +10% of full scale.

The calculation of the input offset is the same as for the 2-point calibration above, while the gain and output offset are calculated for a 200mV full-scale case as:

$$\text{Gain} = \frac{\text{Reading at 200mV} - \text{Reading at 20mV}}{\text{180mV}}$$

$$\text{Output Offset} = \frac{\text{Reading at 200mV} + \text{Reading at} - \text{200mV} - \text{400mV} \bullet \text{Gain}}{2}$$

APPLICATIONS INFORMATION

TROUBLESHOOTING GUIDE

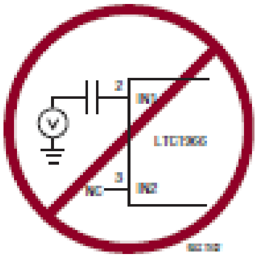
Top Ten LTC1966 Application Mistakes

1. Circuit won't work–Dead On Arrival–no power drawn.
   - Probably forgot to enable the LTC1966 by pulling Pin 8 low.

     Solution: Tie Pin 8 to Pin 1.
2. Circuit won't work, but draws power. Zero or very little output, single-ended input application.
   - Probably didn't connect both input pins.

     Solution: Tie both inputs to something. See Input Connections in the Design Cookbook.

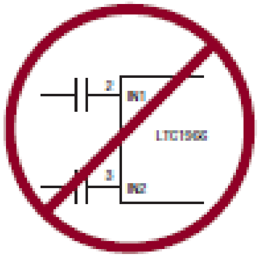
CONNECT PIN 3



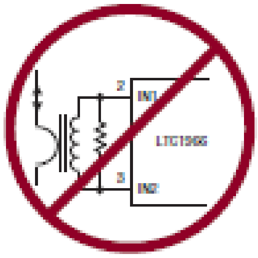
3. Screwy results, particularly with respect to linearity or high crest factors; differential input application.
   - Probably AC-coupled both input pins.

   Solution: Make at least one input DC-coupled. See Input Connections in the Design Cookbook.

DC CONNECT ONE INPUT



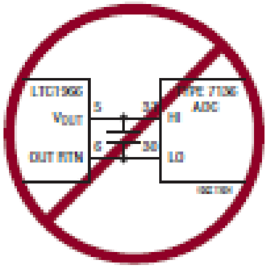
DC CONNECT ONE INPUT



4. Gain is low by a few percent, along with other screwy results.
   - Probably tried to use output in a floating, differential manner.

     Solution: Tie Pin 6 to a low impedance. See Output Connections in the Design Cookbook.

GROUND PIN 6



5. Offsets perceived to be out of specification because 0V in ≠ 0V out.
   - The offsets are not specified at 0V in. No RMS-to-DC converter works well at 0 due to a divide-by-zero calculation.

     Solution: Measure V_{IOS}/V_{OOS} by extrapolating readings > $\pm 5mV_{DC}$.
6. Linearity perceived to be out of specification particularly with small input signals.
   - This could again be due to using 0V in as one of the measurement points.

     Solution: Check Linearity from $5mV_{RMS}$ to $500mV_{RMS}$.
   - The input offset voltage can cause small AC linearity errors at low input amplitudes as well. See Error Analyses section.

     Possible Solution: Include a trim for input offset.

APPLICATIONS INFORMATION

7. Output is noisy with >10kHz inputs.
   - This is a fundamental characteristic of this topology. The LTC1966 is designed to work very well with inputs of 1kHz or less. It works okay as high as 1MHz, but it is limited by aliased ΔΣ noise.

     Solution: Bandwidth limit the input or digitally filter the resulting output.

8. Large errors occur at crest factors approaching, but less than 4.
   - Insufficient averaging.

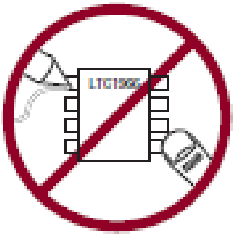
     Solution: Increase C_{AVE}. See Crest Factor and AC + DC Waveforms section for discussion of output droop.

9. Screwy results, errors > spec limits, typically 1% to 5%.
   - High impedance (85kΩ) and high accuracy (0.1%) require clean boards! Flux residue, finger grime, etc. all wreak havoc at this level.

     Solution: Wash the board.

     Helpful Hint: Sensitivity to leakages can be reduced significantly through the use of guard traces.

KEEP BOARD CLEAN

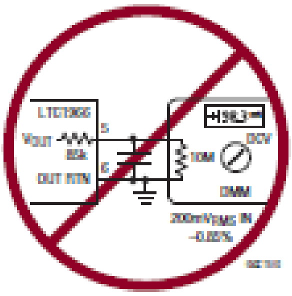


10. Gain is low by ≅1% or more, no other problems.
    - Probably due to circuit loading. With a DMM or a 10× scope probe, Z_{IN} = 10MΩ. The LTC1966 output is 85kΩ, resulting in –0.85% gain error. Output impedance is higher with the DC accurate post filter.

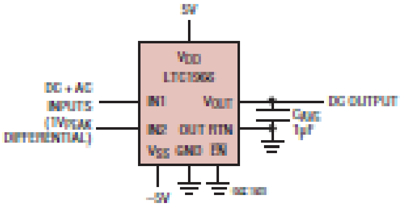
      Solution: Remove the shunt loading or buffer the output.
    - Loading can also be caused by cheap averaging capacitors.

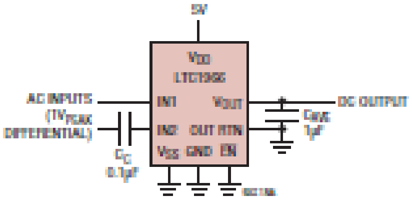
      Solution: Use a high quality metal film capacitor for C_{AVE}.

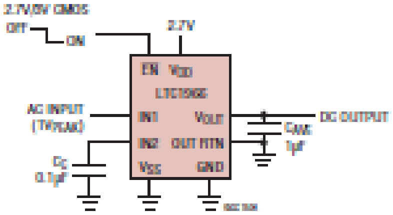
LOADING DRAGS DOWN GAIN

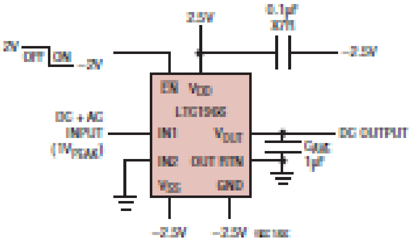


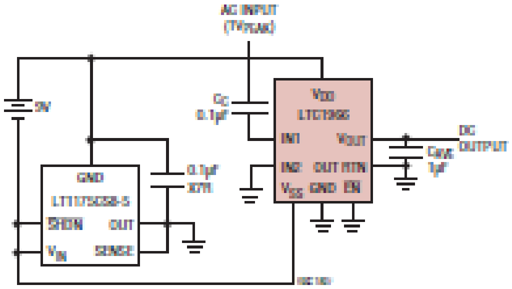
TYPICAL APPLICATIONS


±5V Supplies, Differential, DC-Coupled RMS-to-DC Converter

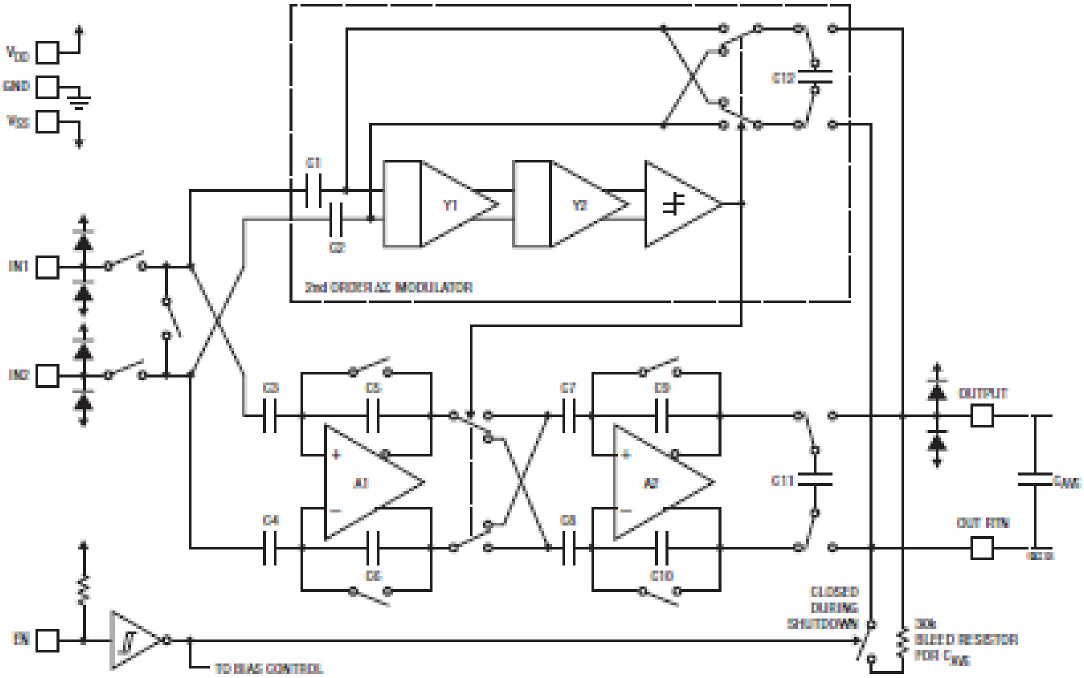

5V Single Supply, Differential, AC-Coupled RMS-to-DC Converter


2.7V Single Supply, Single Ended, AC-Coupled RMS-to-DC Converter with Shutdown


±2.5V Supplies, Single Ended, DC-Coupled RMS-to-DC Converter with Shutdown

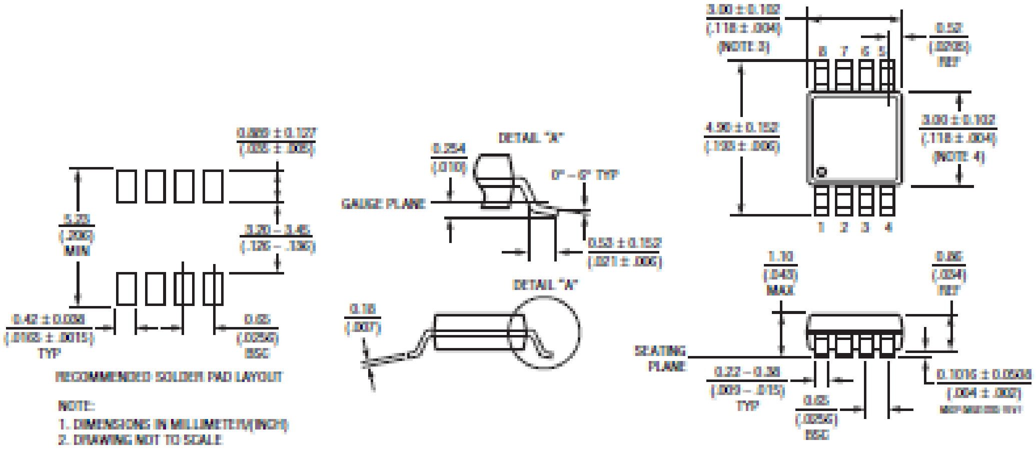

Battery Powered Single-Ended AC-Coupled RMS-to-DC Converter

SIMPLIFIED SCHEMATIC



PACKAGE DESCRIPTION

MS8 Package
8-Lead Plastic MSOP
(Reference LTC DWG # 05-08-1660 Rev F)



RECOMMENDED SOLDER PAD LAYOUT

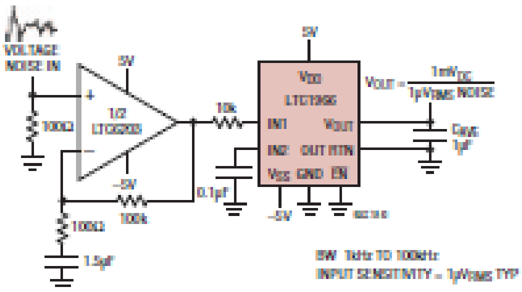
NOTE:
1. DIMENSIONS IN MILLIMETER/(INCH)
2. DRAWING NOT TO SCALE
3. DIMENSION DOES NOT INCLUDE MOLD FLASH, PROTRUSIONS OR GATE BURRS.
   MOLD FLASH, PROTRUSIONS OR GATE BURRS SHALL NOT EXCEED 0.152mm (.006") PER SIDE
4. DIMENSION DOES NOT INCLUDE INTERLEAD FLASH OR PROTRUSIONS.
   INTERLEAD FLASH OR PROTRUSIONS SHALL NOT EXCEED 0.152mm (.006") PER SIDE
5. LEAD COPLANARITY (BOTTOM OF LEADS AFTER FORMING) SHALL BE 0.102mm (.004") MAX

REVISION HISTORY (Revision history begins at Rev B)

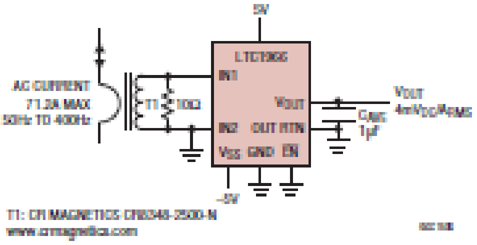
REV	DATE	DESCRIPTION	PAGE NUMBER
B	5/11	Revised entire data sheet to add H- and MP- grades	1 to 38

TYPICAL APPLICATION

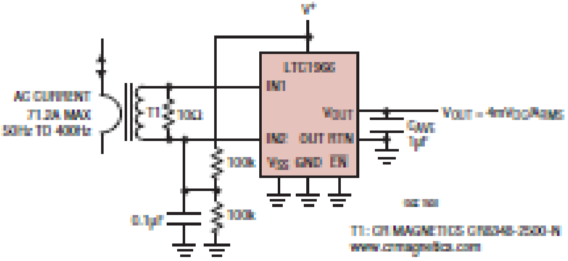
RMS Noise Measurement



70A Current Measurement



Single Supply RMS Current Measurement



RELATED PARTS

PART NUMBER	DESCRIPTION	COMMENTS
LT®1077	Micropower, Single Supply Precision Op Amp	48μA I_{SY}, 60μV $V_{OS(MAX)}$, 450pA $I_{OS(MAX)}$
LT1178-5	Negative, −5V Fixed, Micropower LDO Regulator	45μA I_Q, Available in SO-8 or SOT-223
LT1494	1.5μA Max, Precision Rail-to-Rail I/O Op Amp	375μV $V_{OS(MAX)}$, 100pA $I_{OS(MAX)}$
LT1782	General Purpose SOT-23 Rail-to-Rail Op Amp	40μA I_{SY}, 800μV $V_{OS(MAX)}$, 2nA $I_{OS(MAX)}$
LT1880	SOT-23 Rail-to-Rail Output Precision Op Amp	1.2mA I_{SY}, 150μV $V_{OS(MAX)}$, 900pA $I_{OS(MAX)}$
LTC1967	Precision, Extended Bandwidth RMS to DC Converter	330μA I_{SY}, ΔΣ RMS Conversion to 4MHz
LTC1968	Precision, Wide Bandwidth RMS to DC Converter	2.3mA I_{SY}, ΔΣ RMS Conversion to 15MHz
LTC2050	Zero Drift Op Amp in SOT-23	750μA I_{SY}, 3μV $V_{OS(MAX)}$, 75pA $I_{B(MAX)}$
LT2178/LT2178A	17μA Max, Single Supply Precision Dual Op Amp	14μA I_{SY}, 120μV $V_{OS(MAX)}$, 350pA $I_{OS(MAX)}$
LTC2402	2-Channel, 24-bit, Micropower, No Latency ΔΣ™ ADC	200μA I_{SY}, 4ppm INL, 10ppm TUE
LTC2420	20-bit, Micropower, No Latency ΔΣ ADC in SO-8	200μA I_{SY}, 8ppm INL, 16ppm TUE
LTC2422	2-Channel, 20-bit, Micropower, No Latency ΔΣ ADC	Dual Channel Version of LTC2420

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