

UNIVERSIDADE DE SÃO PAULO
ESCOLA DE ENGENHARIA DE SÃO CARLOS
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LUCAS ALEXANDRE MALAKIN

**VIABILIDADE DE UM SISTEMA DE
ESPECTROFOTOMETRIA PARA AUXÍLIO NO
DIAGNÓSTICO DE ANEMIA**

Trabalho de Conclusão de Curso apresentado
à Escola de
Engenharia de São Carlos, da
Universidade de São Paulo

Curso de Engenharia Elétrica com ênfase em
Eletrônica

ORIENTADOR: Prof. Dr. Carlos Dias Maciel

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AUTORIZO A REPRODUÇÃO TOTAL OU PARCIAL DESTE TRABALHO,
POR QUALQUER MEIO CONVENCIONAL OU ELETRÔNICO, PARA FINS
DE ESTUDO E PESQUISA, DESDE QUE CITADA A FONTE.

M236v Malakin, Lucas Alexandre
Viabilidade de um sistema para auxílio no
diagnóstico de anemia através de espectrofotometria /
Lucas Alexandre Malakin; orientador Carlos Dias Maciel.
São Carlos, 2017.

Monografia (Graduação em Engenharia Elétrica com
ênfase em Eletrônica) -- Escola de Engenharia de São
Carlos da Universidade de São Paulo, 2017.

1. Sangue. 2. Prototipagem. 3. Sistema
opto-eletrônico. I. Título.

FOLHA DE APROVAÇÃO

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Título: "Viabilidade de um sistema para auxílio no diagnóstico de anemia através de espectrofotometria"

Trabalho de Conclusão de Curso defendido e aprovado
em 07 / 11 / 2017,

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AGRADECIMENTOS

Ao meu orientador, professor Carlos Dias Maciel por todo conhecimento compartilhado, pela paciência, pelos conselhos e bom humor.

A todos os meus amigos dos mais diversos círculos sociais, por sempre estarem presentes em todas as situações difíceis. Sobretudo a Andre, Andrei, Arthur, Dani, Guto, Mario, Paulo e Pedro.

Aos meus companheiros de CAASO, por transmitirem o maior conhecimento que se pode obter dentro de uma universidade.

Às famílias que me adotaram, Carneiro, Amaro, Piovezan e Segatto, em cada fase de minha vida e tiveram a paciência de me educar.

Aos meus irmãos Du e Allan, por me aconselharem e acompanharem nessa jornada chamada vida.

A minha namorada, melhor amiga e companheira, Priscila, por trazer a tranquilidade, a sabedoria e maturidade cotidianamente me transformando em um ser humano melhor.

Aos meus pais que, por mais curta que tenha sido a jornada, foram fundamentais para a minha construção.

RESUMO

A anemia é associada com a capacidade do sangue em transportar oxigênio, podendo causar fadiga, cansaço, falta de ar e perda de consciência. O método mais comum para o seu diagnóstico se dá por um hemograma, que envolve a coleta de sangue e tempo para a análise do mesmo. Dessa forma, o diagnóstico pode ser demorado. Com isso, surgiu a ideia do aumento da velocidade do diagnóstico através de um sistema espectrofotométrico com a análise em papel. Dessa forma, estudou-se a viabilidade desse sistema para o diagnóstico dessa condição reduzindo o tempo de diagnóstico e a invasividade do procedimento médico.

Palavra-Chave: Sangue, prototipagem e sistema óptico-eletrônico.

ABSTRACT

Anemia is associated with blood capability to transport oxygen, being able to cause fatigue, weakness, shortness of breath and loss of consciousness. The most common method of diagnosis is by blood count, which involves collecting blood samples and time for analysing it. Thus, the diagnosis can take too long. That is how the idea of increasing the speed of the diagnosis through a spectrophotometric system with analysis in paper arose. Hence, was studied the viability of this system to diagnosis this kind of syndrome, improving not only time of diagnosis but also decreasing the invasiveness of this medical procedure.

Keywords: Blood, prototyping and optoelectronics systems.

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Introdução

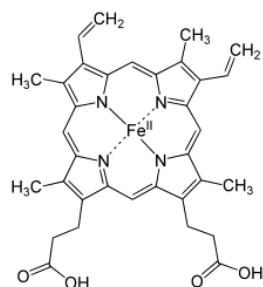
A anemia é uma síndrome na qual a eficiência do sangue em transportar oxigênio para os tecidos está prejudicada/comprometida. Dá-se pela redução no número de hemácias, ou na concentração de hemoglobina no interior dessas células [1]. Com essa redução, tem-se diversas patologias que variam de uma simples sonolência até dispneia.

O método mais comum para o seu diagnóstico é através de um hemograma. Esse teste avalia o número e a forma de células sanguíneas através de uma amostra, coletada do sangue periférico do paciente através de uma seringa e posteriormente transferida para um tubo de ensaio. Trata-se, portanto, de um método invasivo abrupto para a análise [2]. Com isso, tem-se que o procedimento para o diagnóstico demanda uma quantidade de tempo considerável para a análise do sangue.

O sangue humano é um tecido conjuntivo líquido que circula através do sistema vascular. É formado por diversos tipos de células imersas em uma matriz extracelular líquida, o plasma. As células possuem formatos e funções próprias, sendo classificadas em três grandes grupos: os leucócitos ou glóbulos brancos, responsáveis pela defesa do sistema imunitário, as plaquetas, responsáveis pela coagulação sanguínea e os eritrócitos, glóbulos vermelhos ou hemácias, responsáveis pelo transporte de oxigênio.

Como visto, uma redução do transporte de oxigênio pode ser caracterizada pela redução no número de glóbulos vermelhos ou no teor de hemoglobina dessas células [3]. Da deficiência em hemoglobina, tem-se a anemia ferropriva, que é a mais comum dos casos diagnosticados e é associada a baixa concentração de ferro no organismo[4]. O ferro é o principal responsável pelo transporte de oxigênio para os tecidos e fica alojado no núcleo da molécula de hemoglobina, como pode ser visto na Figura 1. Com isso, tem-se que quanto maior a concentração de hemoglobina no organismo, maior será a de ferro.

Figura 1 – Molécula de Hemoglobina



Fonte: LOPES, Sônia¹

A capacidade da hemoglobina de se ligar ao oxigênio depende de uma unidade não peptídica: um grupo heme. Este grupo também dá à hemoglobina a sua cor característica. O grupo heme é constituído por uma parte orgânica e por um átomo de ferro. A parte orgânica, protoporfirina, é constituída por 4 anéis pirrólicos. Os quatro pirróis estão ligados por pontes de metano, para formar um anel tetrapirrólico, como se pode ver na Figura 1.

O átomo de ferro no grupo heme liga-se aos 4 nitrogênios no centro do anel protoporfirínico. O ferro pode formar duas ligações adicionais, uma em cada lado do plano do grupo heme. O átomo de ferro pode estar nos estados de oxidação ferroso (+2) ou férrico (+3), e as formas correspondentes de hemoglobina são denominadas, respectivamente, ferro-hemoglobina e ferri-hemoglobina (também chamada metahemoglobina). Somente a ferro-hemoglobina, o estado de oxidação +2, pode ligar-se ao oxigênio [5]. Dessa forma, uma molécula de hemoglobina é capaz de se combinar com quatro moléculas de gás oxigênio, pois a hemoglobina é tetramérica, ou seja, formada por duas cadeias alfa e duas betas, dessa forma, são quatro grupos heme, os quais cada grupo carrega um grupo oxigênio, assim então, formando a oxiemoglobina.

Com isso, através da espectrofotometria, que é uma técnica que permite comparar a radiação luminosa absorvida ou transmitida por soluções, sendo a concentração de uma dessas conhecida[6][7], visa-se por comparação estimar a concentração da solução desconhecida de ferro e conseqüentemente hemoglobina no sangue e a obtenção de um diagnóstico de anemia.

¹ LOPES, Sônia. Bio 1 ed. São Paulo: Saraiva, 2004

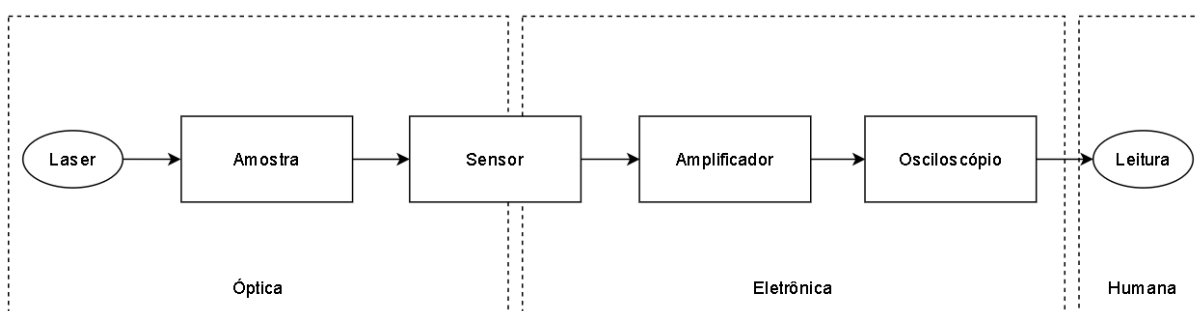
1. Objetivo

Esse trabalho teve por objetivo avaliar a viabilidade de um sistema óptico-eletrônico baseado em um espectrofotômetro para reduzir o tempo de diagnóstico de anemia e a invasividade da coleta de amostra, comparada ao procedimento médico padrão.

2. Materiais e Métodos

O sistema desenvolvido consiste nas divisões óptica e eletrônica e a etapa de testes químicos. A parte óptica se subdivide em iluminação (feita com laser) e detecção do sinal, realizado com sensor óptico, que fornecerá um sinal elétrico de acordo com a luz que o atingir. A etapa eletrônica consiste na fonte de alimentação do circuito associada ao circuito amplificador com o osciloscópio em sua saída. O processo como um todo pode ser visto na Figura 2.

Figura 2: Fluxograma do sistema desenvolvido



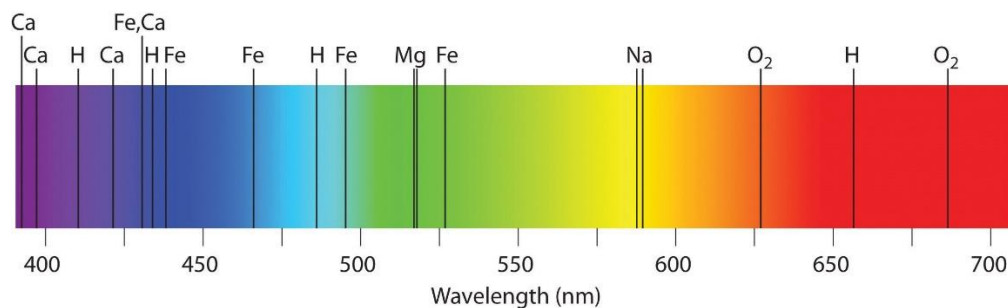
Fonte: Próprio Autor

2.1. Espectrofotometria da Hemoglobina, Ferro e Permanganato de Potássio

Inicialmente, cogitou-se o uso de amostras de sangue para o estudo da viabilidade do sistema para detecção de anemia, no entanto, segundo a Resolução CNS Nº 441, de 12 de maio de 2011, todo uso ou armazenamento de material biológico humano deverá passar pelo conselho de ética. Dessa forma, para contornar esse mecanismo burocrático, buscou-se um composto que possuísse características semelhantes à hemoglobina quanto à banda de absorbância, que não fosse difícil de se obter e possuísse potencial tóxico relativamente baixo.

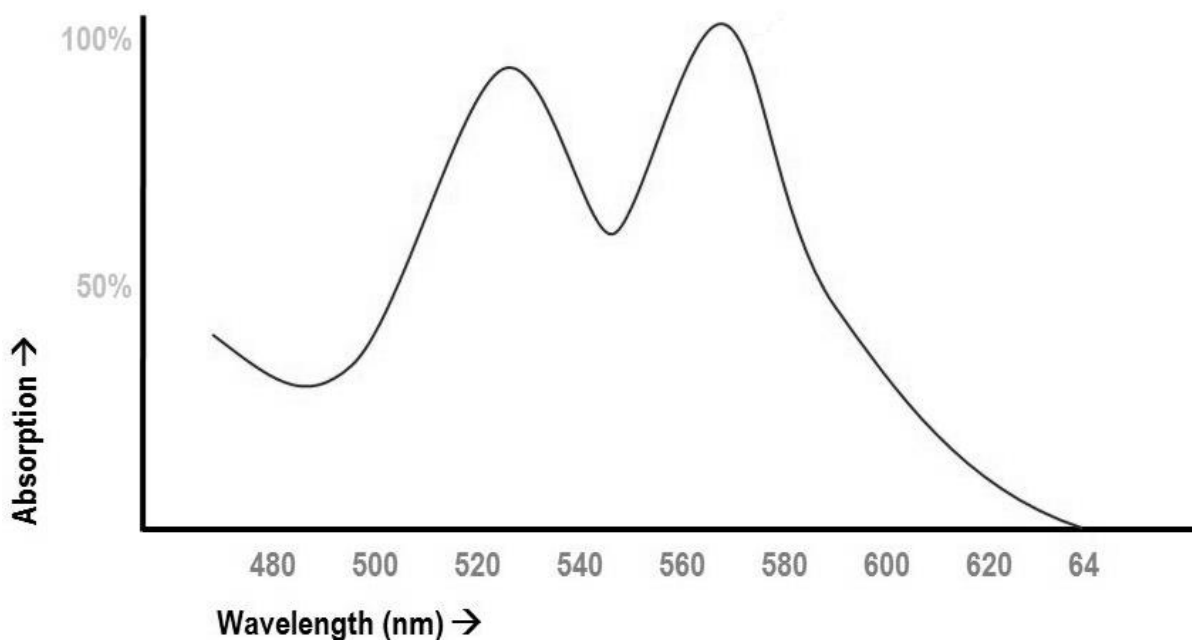
Tendo como base os picos de absorbância por espectrofotometria da hemoglobina, como mostrado na Figura 4, verifica-se que uma boa região para se trabalhar é a de 535nm, que também é onde o ferro (Fe^{2+}) presente no núcleo da molécula de hemoglobina sofre grande interferência da luz, como pode ser visto na Figura 3.

Figura 3: Faixas de frequências e absorbâncias de metais



Fonte: 2012books²

Figura 4 – Espectro de absorbância da hemoglobina



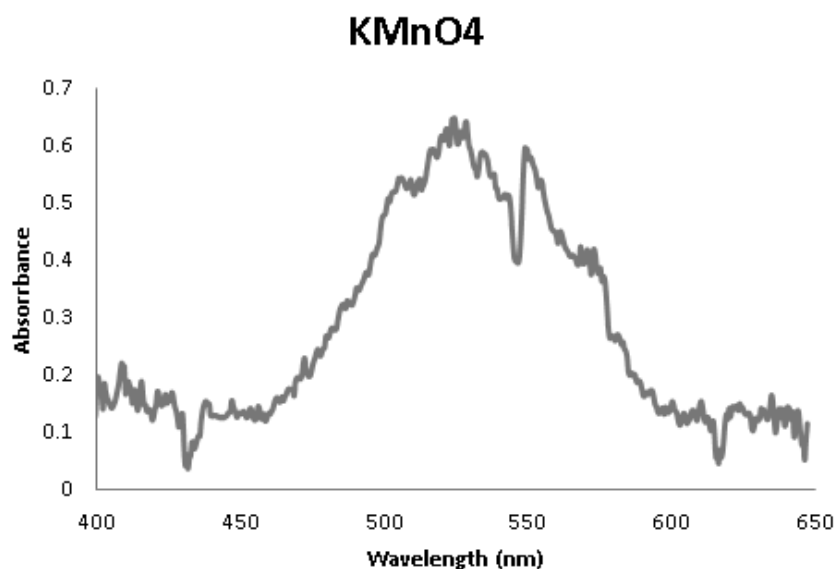
Fonte: Deranged Physiology³

Foi feita uma busca na literatura sobre possíveis compostos para serem utilizados em substituição à amostra biológica (sangue). Foi escolhido, então, o permanganato de potássio (KMnO₄). Na Figura 5, tem-se a curva espectrofotométrica desse composto.

² Disponível em: < <https://2012books.lardbucket.org/books/principles-of-general-chemistry-v1.0/s10-03-atomic-spectra-and-models-of-t.html> > Acesso em out. 2017

³ Disponível em: < <http://www.derangedphysiology.com> > Acesso em out. 2017

Figura 5 – Espectro de absorbância do Permanganato de Potássio



Fonte: Shimadzu⁴

Denota-se que a substância escolhida, KMnO₄, apresenta espectro de absorbância semelhante ao da hemoglobina na frequência de 535nm, sendo, portanto, uma substância adequada para substituir as amostras de sangue nos testes de viabilidade do sistema espectrofotométrico avaliado.

2.2. Óptica

A parte óptica consiste em um sistema de iluminação, feito com laser (marca do laser, modelo, etc). A detecção foi feita de maneira comparativa, mediu-se a tensão no sensor sem qualquer quantidade de permanganato na amostra, e esse valor foi comparado com a tensão obtida nas soluções que continham o sal, de forma a obter-se a absorbância relativa [10]. Com a utilização do laser ao invés de LEDs para o espectro almejado (verde = 535nm), ganha-se muito em qualidade de resultados; isso será demonstrado nas próximas seções dessa monografia.

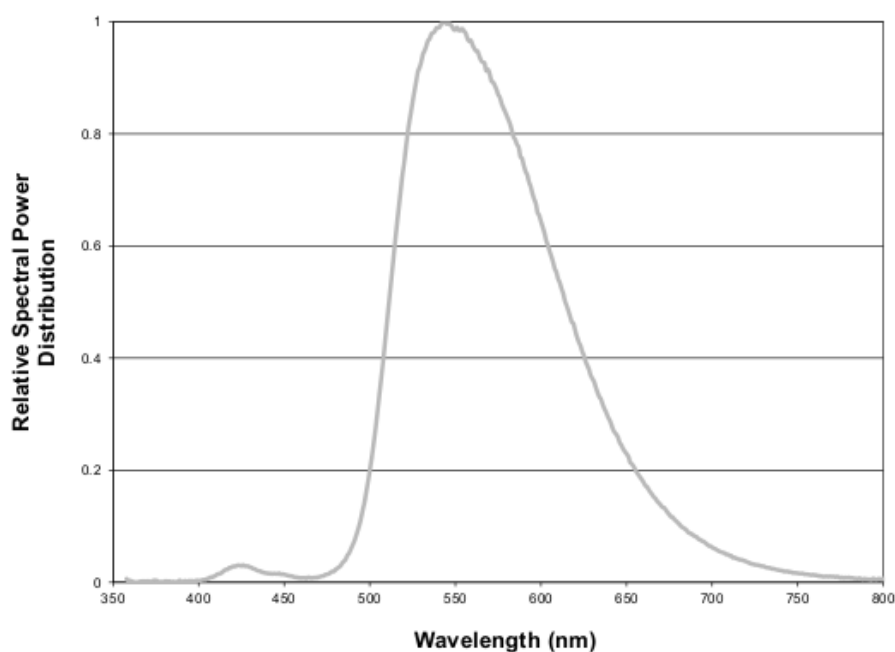
⁴ Disponível em: <<http://www.shimadzu.com/an/uv/support/uv/ap/cell.html>> Acesso em out. 2017

2.3. Iluminação

O tipo de iluminação utilizado é fundamental para a qualidade das medidas, visto que há a necessidade de utilizar um espectro fechado na região de 525nm, pois é a frequência pela qual nossa amostra, permanganato de potássio, sofre efeito de absorção semelhante à hemoglobina. Dessa forma, buscou-se duas fontes populares e baratas de iluminação, que são *LEDs* e Lasers.

O LED é um diodo semicondutor (junção P-N) que quando é energizado emite luz visível – por isso LED (Diodo Emissor de Luz, na sigla em inglês). A luz não é monocromática, mas consiste de uma banda espectral relativamente estreita (exibida na Figura 6 para o LED verde) e é produzida pelas interações energéticas do elétron.

Figura 6 – Espectro de emissão de um LED verde.

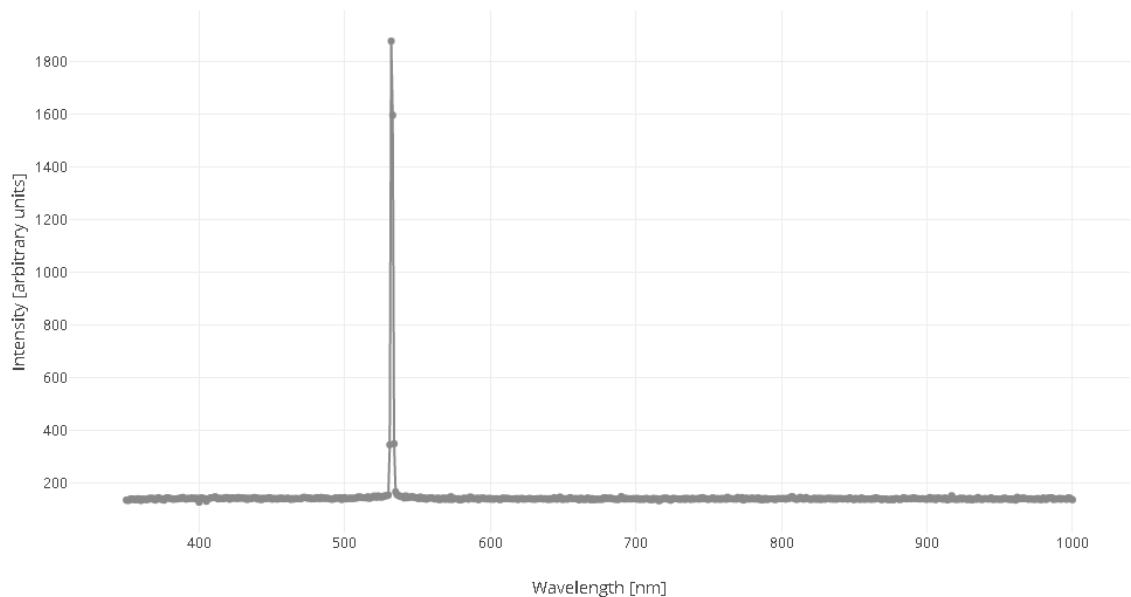


Fonte: Reef2reef⁵

O laser é um dispositivo que produz radiação eletromagnética monocromática, ou seja, possui comprimento de luz extremamente bem definido, além de ser colimada, feixes de luz praticamente paralelos [11]. Na Figura 7, tem-se o espectro de emissão de um laser verde.

⁵ Disponível em: <<https://www.reef2reef.com/threads/mini-evil-cluster-idea-looking-for-feedback.197030/>> Acesso em out. 2017

Figura 7 – Espectro de emissão de um laser verde



Fonte: Photonics⁶

Nota-se que devido a banda espectral ser muito mais fechada do que a do LED, tem-se um ganho elevado na qualidade dos resultados, porque se deseja estudar a frequência nos arredores de 535nm (frequência de máxima absorbância da hemoglobina e do permanganato de potássio).

2.4. Absorbância Relativa

A absorbância é a capacidade intrínseca dos materiais em absorver radiações em uma frequência específica. Usualmente é feita de maneira comparativa, onde se tem a absorbância comparada entre o que é absorvido por um corpo branco e uma amostra específica. Com isso, em espectroscopia a absorbância (A) é dada por:

$$A = -\log_{10} \frac{I}{I_0} \quad (\text{eq. 1})$$

Onde I é a intensidade da luz na amostra específica e I_0 a intensidade da luz na solução cristalina.

⁶ Disponível em: < <https://www.photonics.com/Article.aspx?AID=46771> > Acesso em out. 2017

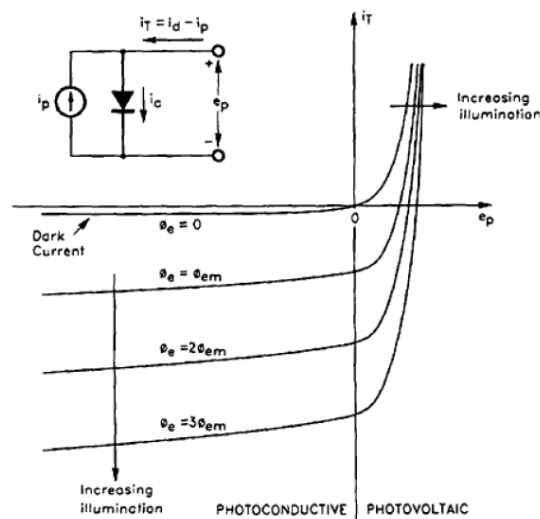
2.5. Fotodiodo

O fotodiodo é um componente eletrônico feito a partir de uma junção semicondutora, que converte a energia de um fóton em um sinal elétrico, liberando e acelerando portadores de condução de carga dentro do semicondutor. Quando um fóton com energia suficiente atinge a junção, ele gera pares elétron-lacuna, levando-os para um estado de condução. Dessa forma, o elétron e a lacuna são varridos pelo campo elétrico na região de depleção da junção e ao se movimentarem geram fotocorrente [12][13].

Sua característica semicondutora é muito semelhante a um diodo comum, ou seja, possui dois modos de utilização destaques, que são fotovoltaico e fotocondutivo. Em ambos os modos, a fotocorrente varia linearmente com a intensidade da luz, no entanto, no modo fotovoltaico as variações da fonte luminosa variam a tensão na junção de maneira logarítmica.

No modo fotocondutivo, há a presença de uma corrente de polarização denotada como “*dark current*”, oriunda da resistência interna da junção que está polarizada. Com isso, tem-se um *offset* na fotocorrente que pode representar um erro nas medidas quando ocorre o cessar da fonte luminosa. Tem-se na Figura 8 a diferença entre os modos fotovoltaico e fotocondutivo.

Figura 8 – Curva que demonstra a diferença entre os modos fotocondutivo e fotovoltaico.



Fonte: GRAEME, J.G.⁷

⁷ GRAEME, J.G. Applications of operational Amplifiers. McGraw-Hill Kogakusha Ltda., 1973. STOUT & KAUFMAN Handbook of operational amplifiers. McGraw-Hill, 1976.

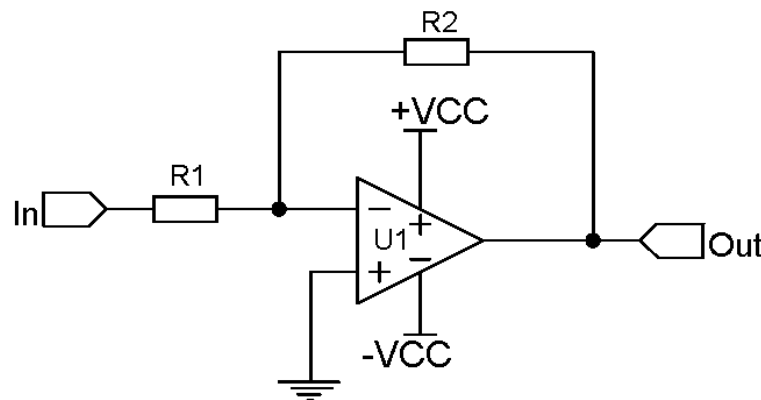
2.6. Eletrônica

A parte eletrônica consiste de um sensor (fotodiodo já exibido no tópico anterior), um circuito amplificador e um sistema para coletar as medidas que pode ser desde um osciloscópio até um microcontrolador. Utilizou-se um osciloscópio para ter um ganho no tempo das análises das amostras, pois se trata de um protótipo, já o fotodiodo escolhido foi o BPW34 (consta no apêndice dessa monografia a folha de dados do dispositivo).

2.7. Circuito amplificador

O amplificador utilizado para o experimento foi o inversor, pois permite um alto ganho, uma impedância de entrada elevada e uma configuração relativamente simples e com poucos dispositivos[12]. Na figura 9 está representado o circuito básico.

Figura 9 – Amplificador inversor



Fonte: GRAEME, J.G.⁷

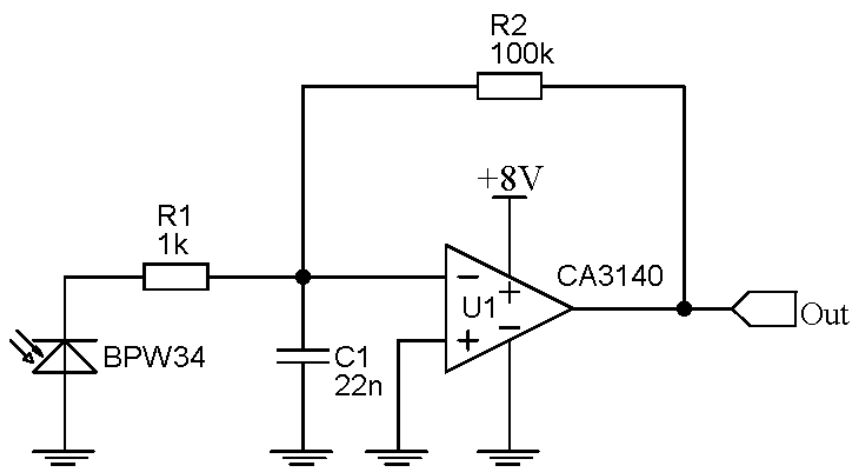
Sua função de transferência é dada por:

$$H = \frac{V_{out}}{V_{in}} = -\frac{R2}{R1} \quad (\text{eq. 2})$$

Nota-se que um único resistor de realimentação, R2, atua como o ganho da fotocorrente e gera com ela a transformação para a tensão de saída. Com isso, deve-se evitar resistores de valores altos, pois pode ocasionar um grande impacto a presença de ruído térmico ao nosso sinal, caso ocorra a utilização de resistores de ordem elevada, pode-se

utilizar uma rede *tee* [12]. Na Figura 10, tem-se o circuito calculado para a coleta e análise do sinal.

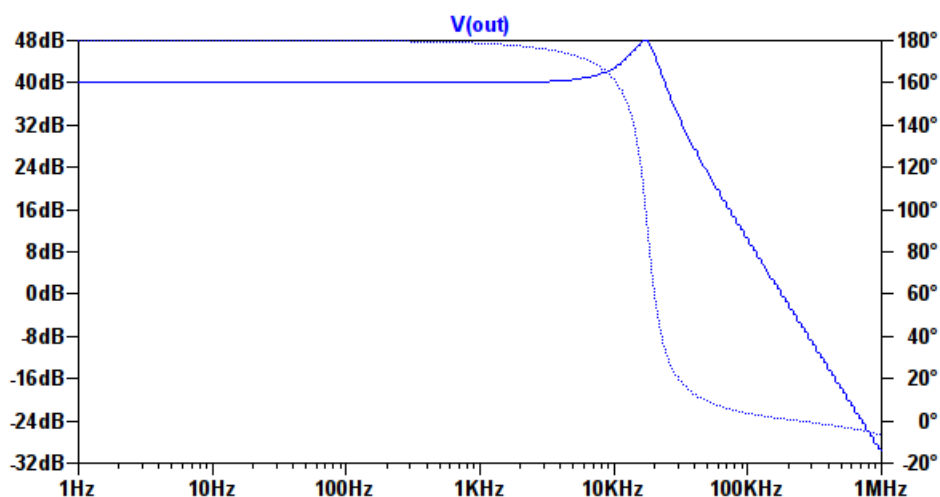
Figura 10 – Configuração do amplificador com o fotodiodo



Fonte: O próprio autor

O uso do amplificador operacional CA3140 se dá pela alta impedância de entrada ($1,5T\Omega$) e pela recomendação do dispositivo em aplicações opto eletrônicas [12]. Nota-se que houve a adição de um capacitor de 22nF a entrada do circuito, isso se dá para filtrar os sinais indesejados de alta frequência, e um ganho de 100 vezes seguindo a equação 2 (40dB). Na Figura 11, tem-se o ganho do circuito para pequenos sinais.

Figura 11 – Ganho do circuito amplificador

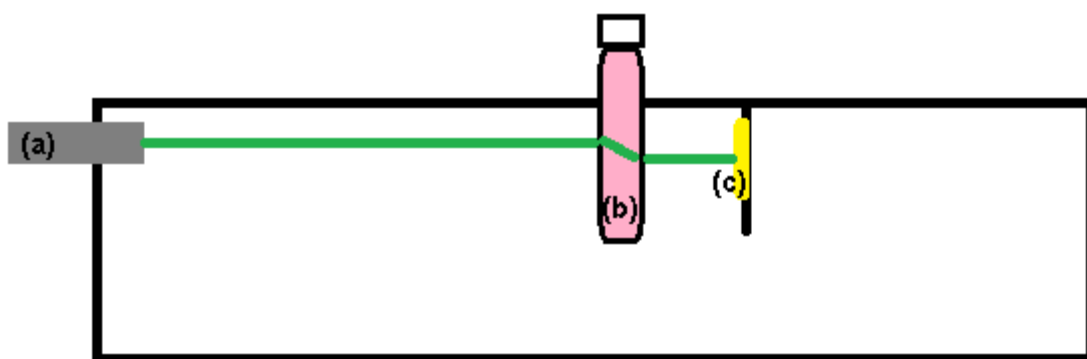


Fonte: O próprio autor

2.8. Corpo do Protótipo

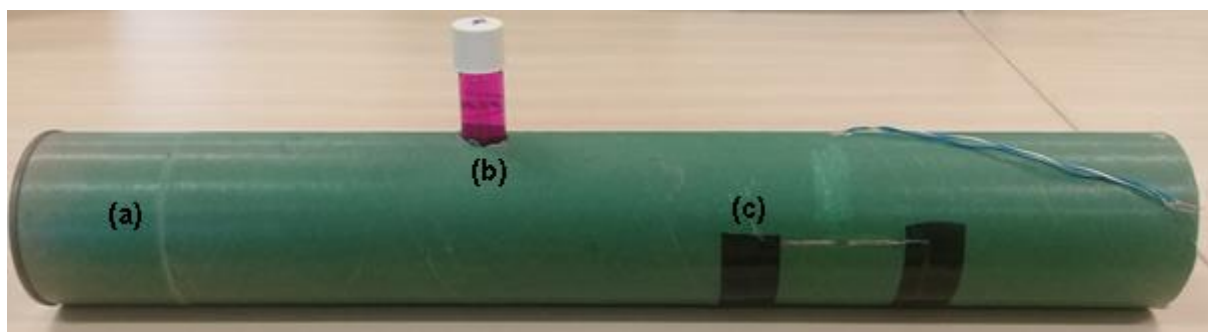
Do fluxograma da Figura 3, elaborou-se com materiais de fácil acesso o protótipo da Figura 12 e o real na Figura 13.

Figura 12 – Protótipo idealizado



Fonte: O próprio autor

Figura 13 – Corpo real do protótipo



Fonte: O próprio autor

Nas Figuras 13 e 14, tem-se em (a) o laser, em (b) a amostra e em (c) nosso fotodiodo.

3. Resultados

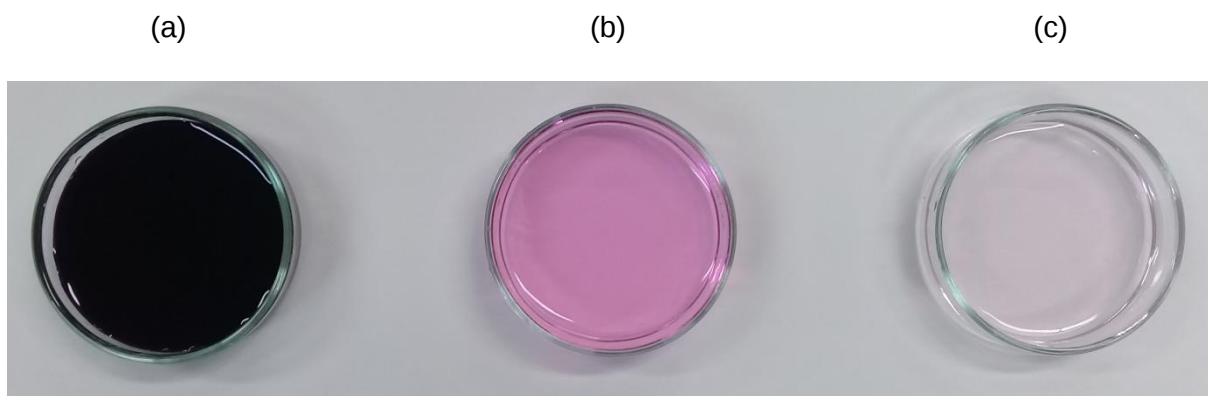
Por se tratar de um experimento que avalia a viabilidade de um sistema que utiliza de distintas áreas, tais como química e optoeletrônica, todas as etapas foram cautelosamente preparadas para garantir a saúde dos testes bem como os resultados. Dessa forma, os resultados serão segmentados em dois tópicos, esses citados acima.

3.1. Preparo das Soluções

O preparo das soluções de permanganato de potássio envolveu a necessidade de cautela devido a riscos que o composto poderia causar. Segundo a ficha de informação de produto químico[15], o uso de equipamentos de proteção individual (EPI) se faz necessário devido ao sal inorgânico ser levemente tóxico, podendo causar irritação na pele, olhos e se caso ingerido, náusea, vômito ou perda da consciência. Além disso, em contato com superfícies o KMnO_4 mancha, o que ocorreu durante o experimento.

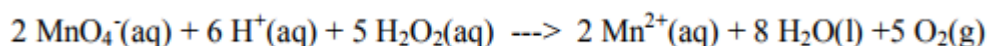
Para a remoção dessas manchas que ocorrem com uma solução semelhante da figura 14(a), deve-se aplicar vinagre que resultada na solução da figura 14(b). Posterior a isso, deve-se aplicar água oxigenada, dessa forma, o produto resultado é incolor como a solução da figura 14(c).

Figura 14: Soluções de Permanganato de Potássio



Fonte: O próprio autor

Quimicamente, tem-se a seguinte reação:



Nessa reação, tem-se a redução do íon permanganato (MnO_4^-). A solução desse íon apresenta uma coloração violeta figura 14(a) e quando reage com vinagre e água oxigenada forma o íon manganês, dessa forma, torna-se incolor figura 14(c) [16]. A balança de precisão utilizada pode ser vista na Figura 15.

Figura 15: Balança de precisão utilizada



Fonte: O próprio autor

Esse instrumento de medição possui precisão de 0,001g e máxima medida de 340g.

Optou-se por trabalhar com diluições, dessa forma, prepara-se uma única solução direta com o manuseio de permanganato e adiciona-se água para se obter solução com concentrações distintas desse sal. Com isso, afim de obter a absorbância relativa, tem-se uma solução padrão (água) e outras quatro obtidas a partir da diluição de uma primeira.

Diluição é o ato físico-químico de tornar uma solução menos concentrada em partículas de soluto através do aumento do solvente (água) [17][18]. A solução 1 foi obtida pela dissolução do sal ($\text{KMnO}_4 = 1,200\text{g}$) em água (153,548 ml) e sua concentração (C_{S2}) é calculada pela equação 3.

$$C = \frac{m}{V} \quad (\text{eq.3})$$

Onde m é a massa do soluto e V o volume do solvente.

$$C_{S2} = \frac{m2}{V2} = \frac{1,2}{153,548} = 8,211 \cdot 10^{-3} \text{ g/ml}$$

A solução 3 foi obtida a partir da diluição da solução 2, e sua concentração (C_{S3}) é calculada a partir da equação 4.

$$C_a \cdot V_a = C_b \cdot V_b \quad (\text{eq. 4})$$

Onde C_a e V_a são respectivamente concentração e volume de uma solução A e C_b e V_b são respectivamente concentração e volume de uma solução B.

$$C_{S3} \cdot V3 = C_{S2} \cdot V2$$

$$C_{S3} = \frac{C_{S2} \cdot V2}{V3} = \frac{8,211 \cdot 10^{-3} \cdot 69,796}{160,442} = 3,572 \cdot 10^{-3} \text{ g/ml}$$

A solução 4 foi obtida a partir da diluição da solução 3, e sua concentração (C_{S4}) é calculada a partir da equação 4.

$$C_{S4} \cdot V4 = C_{S3} \cdot V3$$

$$C_{S4} = \frac{C_{S3} \cdot V3}{V4} = \frac{3,572 \cdot 10^{-3} \cdot 65,976}{179,496} = 1,313 \cdot 10^{-3} \text{ g/ml}$$

A solução 5 foi obtida a partir da diluição da solução 4, e sua concentração (C_{S5}) é calculada a partir da equação 4.

$$C_{S5} \cdot V5 = C_{S4} \cdot V4$$

$$C_{S5} = \frac{C_{S4} \cdot V4}{V5} = \frac{1,313 \cdot 10^{-3} \cdot 57,711}{192,033} = 0,395 \cdot 10^{-3} \text{ g/ml}$$

Um resumo das soluções pode ser visto na Tabela 1 e as amostras na Figura 16.

Tabela 1 – Concentrações das soluções

Solução	[KMnO ₄] g/ml
1	----
2	$8,211 \cdot 10^{-3}$
3	$3,572 \cdot 10^{-3}$
4	$1,313 \cdot 10^{-3}$
5	$0,395 \cdot 10^{-3}$

Figura 16: Amostras preparadas



Fonte: O próprio autor

3.2. Experimento Opto-eletrônico

Com as amostras prontas, deu-se início à coleta dos valores absorvidos pelas mesmas. Considerando encontrar valores para *threshold* mínimo (sem qualquer fonte de luz) e máximo (solução sem sal), mediu-se as tensões de saída (em *out*) do circuito da Figura 11.

Tabela 2 – Tensão de saída do circuito para cada amostra

Saída (mV)	
Laser Desligado	120
Solução sem permanganato	1020
Amostra 2	606
Amostra 3	716
Amostra 4	784
Amostra 5	846

Dessa forma, pode-se calcular a absorvância relativa (na equação 5) da amostra. A comparação será feita com o branco (solução sem permanganato) e garante que só será avaliada a absorvância relativa ao soluto de interesse (permanganato de potássio) e a absorvância do solvente e as perdas por reflexões serão desconsideradas[19].

$$A = -\log\left(\frac{I_{\text{amostra}}}{I_{\text{branco}}}\right) \quad (\text{eq.5})$$

Com isso, desconsiderando o laser desligado e assumindo o I_{branco} para cada medida da coleta da solução sem permanganato, tem-se a tabela 3 que demonstra a absorvância relativa das amostras.

Tabela 3 – Absorvância relativa das amostras

Absorvância Relativa	
Solução sem permanganato	0,0000*
Amostra 2	0,2261
Amostra 3	0,1537
Amostra 4	0,1143
Amostra 5	0,0812

3.3. Correlações

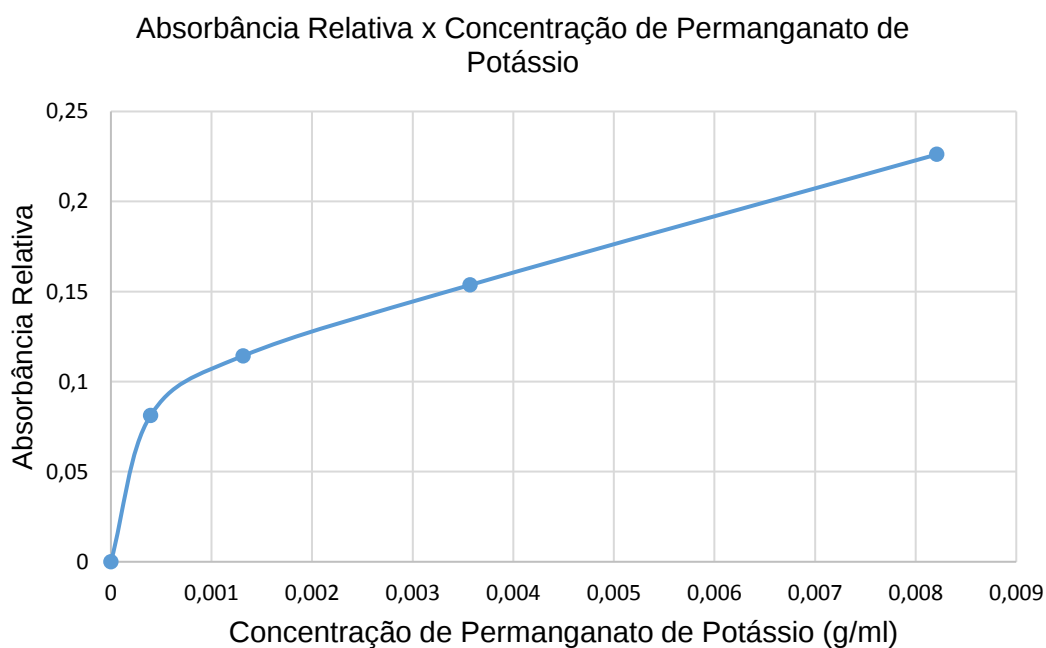
Correlacionando as informações das tabelas 1 e 3, tem-se uma análise da absorbância pela concentração utilizada em ordem crescente da concentração de KMnO_4 , mostradas na tabela 4.

Tabela 4 – Absorbância para cada concentração de permanganato de potássio.

	[KMnO_4] (g/ml)	Absorbância
Solução sem permanganato	$0,000 \cdot 10^{-3}$	0,02648
Amostra 2	$8,211 \cdot 10^{-3}$	0,2261
Amostra 3	$3,572 \cdot 10^{-3}$	0,1537
Amostra 4	$1,313 \cdot 10^{-3}$	0,1143
Amostra 5	$0,395 \cdot 10^{-3}$	0,0812

Da tabela 4, pode-se gerar o gráfico da Figura 17.

Figura 17 – Gráfico das absorbâncias relativas em função das concentrações



Fonte: O próprio autor.

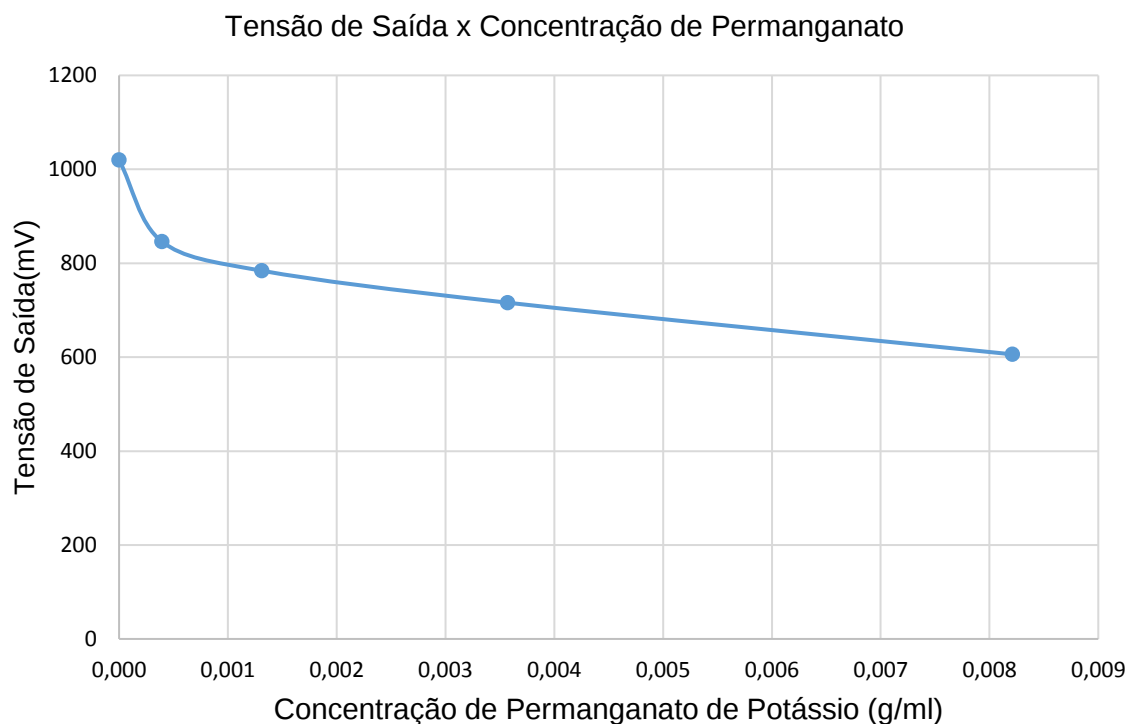
Nota-se que há também a possibilidade da construção de dados a partir da correlação das tabelas 1 e 2, dessa forma, tem-se a tabela 5 que correlaciona as concentrações das amostras com as tensões de saída.

Tabela 5 – Tensão de saída relacionada com a concentração de cada amostra

	[KMnO ₄] (g/ml)	V(out) (mV)
Solução sem permanganato	0,000	1020
Amostra 2	$8,211 \cdot 10^{-3}$	606
Amostra 3	$3,572 \cdot 10^{-3}$	716
Amostra 4	$1,313 \cdot 10^{-3}$	784
Amostra 5	$0,395 \cdot 10^{-3}$	846

Da tabela 5, pode-se gerar o gráfico da Figura 18.

Figura 18 – Gráfico das tensões de saída para cada concentração de amostra testada.



Fonte: O próprio autor.

Na Figura 17, tem-se que quanto maior a concentração de permanganato de potássio maior será a absorbância da amostra, com isso, se a amostra absorve mais luz, quanto maior a concentração de soluto (KMnO_4) menor será a tensão de saída, o que pode ser visto na Figura 18.

4. Conclusão

O estudo da viabilidade de um sistema espectrofotométrico para diagnóstico de anemia se demonstrou promissor. Com um diagnóstico em tempo real, mesmo que em permanganato de potássio e não em amostras sanguíneas, obteve-se curvas muito boas com características logarítmicas, como é de se esperar ao se tratar de absorbância e transmitância.

Do experimento em si há de se destacar o preparo das soluções, pois houve a necessidade de reaprender a manusear algumas vidrarias, balança de precisão e procedimentos de segurança. Mesmo que com um acidente envolvendo o derramamento da solução na bancada, aprendeu-se maneira de remover um sal com alto potencial oxidante de maneira eficaz e segura.

Com os resultados obtidos do protótipo, há de concluir que o sistema é eficaz, pois ocorre a variação da tensão de saída de maneira que se varia as concentrações das amostras para uma mesma fonte luminosa, laser com frequência de 535nm, entretanto, como melhoria contínua há a necessidade de um estudo envolvendo material biológico, humano ou não, para se obterem resultados com condições reais de diagnóstico.

Como melhoria futura, ainda se deve considerar a utilização de uma banda de frequência luminosa maior, pois com isso há a possibilidade de obter a concentração de outras substâncias, com isso, em um estudo mais prolongado e com a estruturação de um banco de dados, pode-se utilizar de redes bayesianas para fazer diagnóstico de outras síndromes, doenças e etc.

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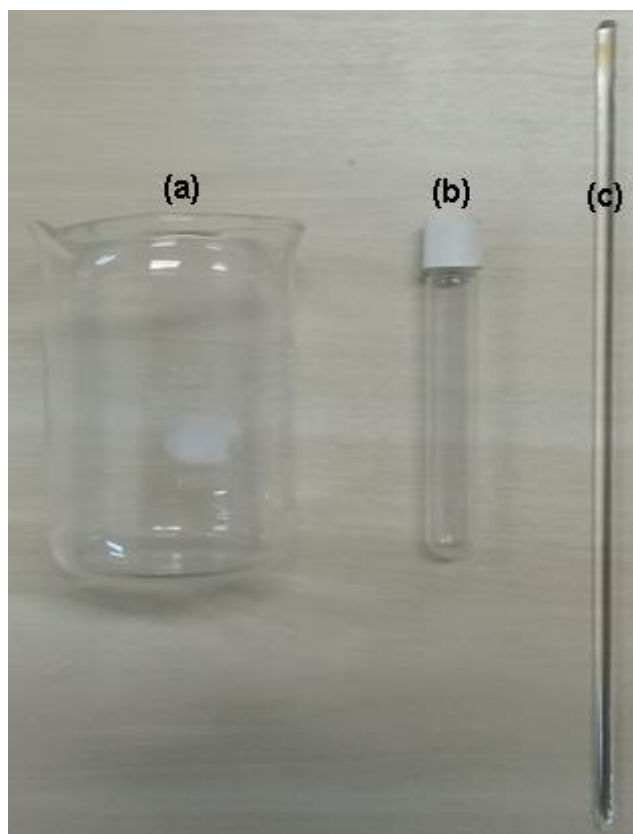
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Apêndice I: Vidraria

Na Figura 19, tem-se em (a) um béquer, utilizado para preparar as soluções, em (b) um tubo de ensaio com tampa de rosca, utilizado para armazenar a solução e em (c) um bastão de vidro utilizado para misturar o soluto com solvente no béquer.

Figura 19: Vidraria utilizada



Fonte: O próprio autor

O béquer deve ser manuseado com total segurança para não ocorrer derramamento da solução dentro da balança. Nota-se que é aconselhável o uso de um papel que possa absorver eventuais gotas da solução entre a base desse dispositivo e a balança.

Faz-se aconselhável o uso de um funil para encher o tubo de ensaio a partir do béquer.

No processo de mistura realizado com o bastão de vidro, deve-se tomar cuidado para não ocorrer acidente com qualquer porção da solução.

No processo todo, aconselha-se o uso de luvas de borracha e proteção da bancada para que não ocorra manchas acidentais pelo derramamento da solução.

Caso ocorra alguma mancha, utiliza-se de vinagre e posteriormente de água oxigenada para a remoção.

Apêndice II: Roteiro químico

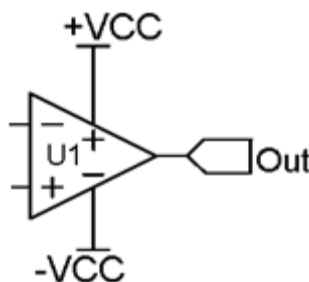
1. Mede-se a massa do papel não absorvente a qual será disposto o permanganato de potássio.
2. Deposita-se a quantidade do sal sobre o papel até se obter a massa almejada.
3. A diferença entre os passos 1 e 2 é a massa do sal.
4. Mede-se a massa do béquer vazio.
5. Preenche-se o béquer de água até atingir o volume almejado.
6. A diferença entre os passos 5 e 4 é a massa de água utilizada. Nota-se que a densidade da água é aproximadamente 1g/ml.
7. Adicionando o permanganato a água, tem-se uma solução que tem a concentração obtida pela eq.3.
8. Descarta-se a quantidade desejada da solução do passo 7, mede-se a massa.
9. Adiciona-se água até obter por dissolução a concentração desejada, nota-se que pode ser obtida essa nova concentração de acordo com a eq.4.
10. Repete-se os passos 8 e 9 a quantidade de vezes suficientes que se deseja obter a solução diluída.

Apêndice III: O amplificador operacional

O amplificador operacional (amp op) é um amplificador que pode ter ganho muito elevado, possui dois terminais de entrada (um terminal não inversor e um inversor) e a sua tensão de saída é a diferença entre os terminais citados acima, multiplicado pelo ganho em malha aberta[12].

O amp op ideal tem ganho, largura de banda e impedância infinitos e ruído, offset de entrada e impedância de saída nulos. Pode ser empregado nas mais diversas aplicações dentro dos ramos da eletrônica e possui uma infinidade de configurações[12].

Figura 20: Amplificador Operacional Básico



Fonte: O próprio autor

Anexos

Anexo A: Folha de dados do fotodiodo BPW34



www.vishay.com

BPW34, BPW34S

Vishay Semiconductors

Silicon PIN Photodiode



94 8583

FEATURES

- Package type: leaded
- Package form: top view
- Dimensions (L x W x H in mm): 5.4 x 4.3 x 3.2
- Radiant sensitive area (in mm²): 7.5
- High photo sensitivity
- High radiant sensitivity
- Suitable for visible and near infrared radiation
- Fast response times
- Angle of half sensitivity: $\phi = \pm 65^\circ$
- Compliant to RoHS Directive 2002/95/EC and in accordance to WEEE 2002/96/EC



RoHS
COMPLIANT
GREEN
(S-2008)**

Note

** Please see document "Vishay Material Category Policy":
www.vishay.com/doc?99902

APPLICATIONS

- High speed photo detector

DESCRIPTION

BPW34 is a PIN photodiode with high speed and high radiant sensitivity in miniature, flat, top view, clear plastic package. It is sensitive to visible and near infrared radiation. BPW34S is packed in tubes, specifications like BPW34.

PRODUCT SUMMARY

COMPONENT	I_{ra} (μA)	ϕ (deg)	$\lambda_{0.1}$ (nm)
BPW34	50	± 65	430 to 1100
BPW34S	50	± 65	430 to 1100

Note

- Test condition see table "Basic Characteristics"

ORDERING INFORMATION

ORDERING CODE	PACKAGING	REMARKS	PACKAGE FORM
BPW34	Bulk	MOQ: 3000 pcs, 3000 pcs/bulk	Top view
BPW34S	Tube	MOQ: 1800 pcs, 45 pcs/tube	Top view

Note

- MOQ: minimum order quantity

ABSOLUTE MAXIMUM RATINGS ($T_{amb} = 25^\circ C$, unless otherwise specified)

PARAMETER	TEST CONDITION	SYMBOL	VALUE	UNIT
Reverse voltage		V_R	60	V
Power dissipation	$T_{amb} \leq 25^\circ C$	P_V	215	mW
Junction temperature		T_J	100	$^\circ C$
Operating temperature range		T_{amb}	- 40 to + 100	$^\circ C$
Storage temperature range		T_{stg}	- 40 to + 100	$^\circ C$
Soldering temperature	$t \leq 3$ s	T_{sd}	260	$^\circ C$
Thermal resistance junction/ambient	Connected with Cu wire, 0.14 mm ²	R_{thJA}	350	K/W

Rev. 2.1, 23-Aug-11

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Document Number: 81521

For technical questions, contact: detectorsupport@vishay.com

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**BASIC CHARACTERISTICS** ($T_{amb} = 25\text{ }^{\circ}\text{C}$, unless otherwise specified)

PARAMETER	TEST CONDITION	SYMBOL	MIN.	TYP.	MAX.	UNIT
Breakdown voltage	$I_R = 100\text{ }\mu\text{A}$, $E = 0$	$V_{(BR)}$	60			V
Reverse dark current	$V_R = 10\text{ V}$, $E = 0$	I_{ro}		2	30	nA
Diode capacitance	$V_R = 0\text{ V}$, $f = 1\text{ MHz}$, $E = 0$	C_D		70		pF
	$V_R = 3\text{ V}$, $f = 1\text{ MHz}$, $E = 0$	C_D		25	40	pF
Open circuit voltage	$E_e = 1\text{ mW/cm}^2$, $\lambda = 950\text{ nm}$	V_o		350		mV
Temperature coefficient of V_o	$E_e = 1\text{ mW/cm}^2$, $\lambda = 950\text{ nm}$	TK_{V_o}		- 2.6		mV/K
Short circuit current	$E_A = 1\text{ klx}$	I_k		70		μA
	$E_e = 1\text{ mW/cm}^2$, $\lambda = 950\text{ nm}$	I_k		47		μA
Temperature coefficient of I_k	$E_e = 1\text{ mW/cm}^2$, $\lambda = 950\text{ nm}$	TK_{I_k}		0.1		%/K
Reverse light current	$E_A = 1\text{ klx}$, $V_R = 5\text{ V}$	I_{ra}		75		μA
	$E_e = 1\text{ mW/cm}^2$, $\lambda = 950\text{ nm}$, $V_R = 5\text{ V}$	I_{ra}	40	50		μA
Angle of half sensitivity		ϕ		± 65		deg
Wavelength of peak sensitivity		λ_p		900		nm
Range of spectral bandwidth		$\lambda_{0.1}$		430 to 1100		nm
Noise equivalent power	$V_R = 10\text{ V}$, $\lambda = 950\text{ nm}$	NEP		4×10^{-14}		W/ $\sqrt{\text{Hz}}$
Rise time	$V_R = 10\text{ V}$, $R_L = 1\text{ k}\Omega$, $\lambda = 820\text{ nm}$	t_r		100		ns
Fall time	$V_R = 10\text{ V}$, $R_L = 1\text{ k}\Omega$, $\lambda = 820\text{ nm}$	t_f		100		ns

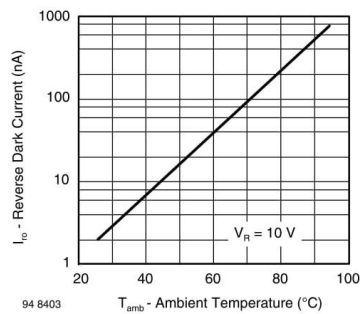
BASIC CHARACTERISTICS ($T_{amb} = 25\text{ }^{\circ}\text{C}$, unless otherwise specified)

Fig. 1 - Reverse Dark Current vs. Ambient Temperature

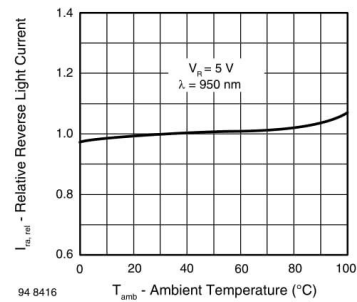


Fig. 2 - Relative Reverse Light Current vs. Ambient Temperature

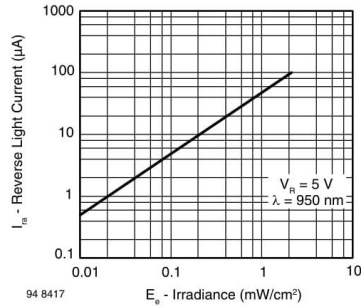


Fig. 3 - Reverse Light Current vs. Irradiance

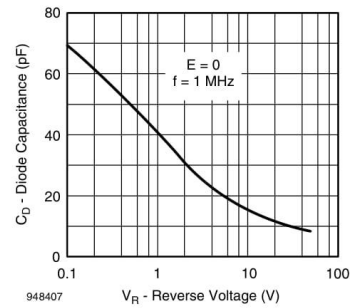


Fig. 6 - Diode Capacitance vs. Reverse Voltage

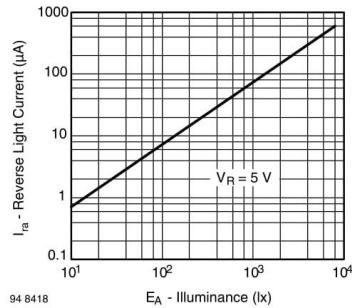


Fig. 4 - Reverse Light Current vs. Illuminance

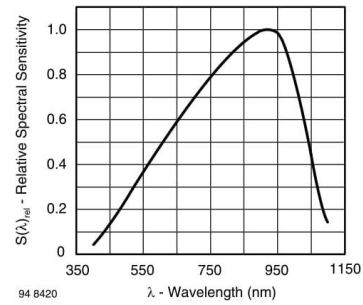


Fig. 7 - Relative Spectral Sensitivity vs. Wavelength

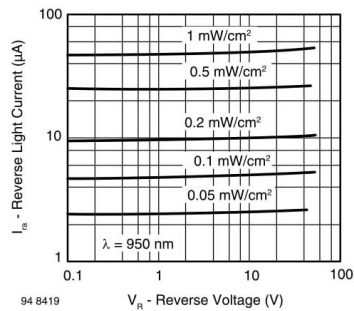


Fig. 5 - Reverse Light Current vs. Reverse Voltage

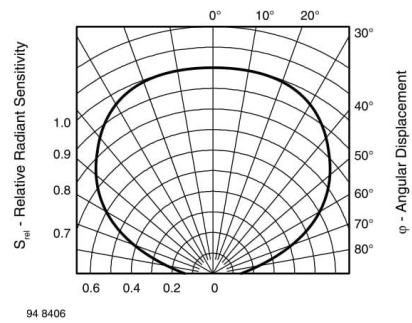
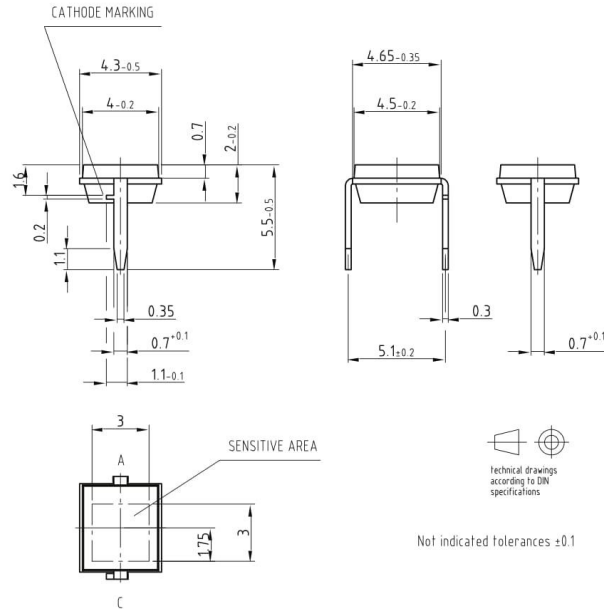


Fig. 8 - Relative Radiant Sensitivity vs. Angular Displacement



PACKAGE DIMENSIONS in millimeters



Drawing-No.: 6.544-5315.01-4
Issue: 1, 19.10.07
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TUBE PACKAGING DIMENSIONS in millimeters

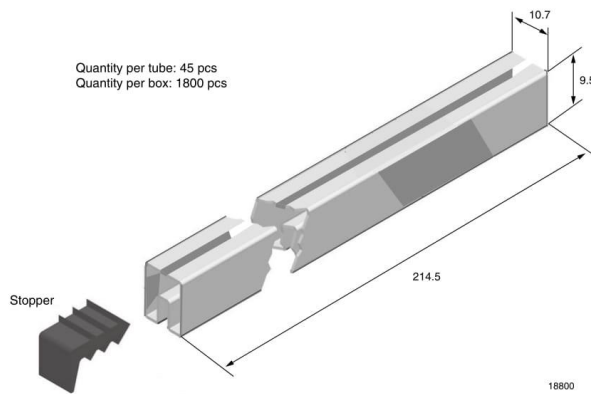


Fig. 9 - Drawing Proportions not scaled



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CA3140, CA3140A

Data Sheet

September 1998

File Number 957.4

4.5MHz, BiMOS Operational Amplifier with MOSFET Input/Bipolar Output

The CA3140A and CA3140 are integrated circuit operational amplifiers that combine the advantages of high voltage PMOS transistors with high voltage bipolar transistors on a single monolithic chip.

The CA3140A and CA3140 BiMOS operational amplifiers feature gate protected MOSFET (PMOS) transistors in the input circuit to provide very high input impedance, very low input current, and high speed performance. The CA3140A and CA3140 operate at supply voltage from 4V to 36V (either single or dual supply). These operational amplifiers are internally phase compensated to achieve stable operation in unity gain follower operation, and additionally, have access terminal for a supplementary external capacitor if additional frequency roll-off is desired. Terminals are also provided for use in applications requiring input offset voltage nulling. The use of PMOS field effect transistors in the input stage results in common mode input voltage capability down to 0.5V below the negative supply terminal, an important attribute for single supply applications. The output stage uses bipolar transistors and includes built-in protection against damage from load terminal short circuiting to either supply rail or to ground.

The CA3140 Series has the same 8-lead pinout used for the "741" and other industry standard op amps. The CA3140A and CA3140 are intended for operation at supply voltages up to 36V ($\pm 18V$).

Ordering Information

PART NUMBER (BRAND)	TEMP. RANGE (°C)	PACKAGE	PKG. NO.
CA3140AE	-55 to 125	8 Ld PDIP	E8.3
CA3140AM (3140A)	-55 to 125	8 Ld SOIC	M8.15
CA3140AS	-55 to 125	8 Pin Metal Can	T8.C
CA3140AT	-55 to 125	8 Pin Metal Can	T8.C
CA3140E	-55 to 125	8 Ld PDIP	E8.3
CA3140M (3140)	-55 to 125	8 Ld SOIC	M8.15
CA3140M96 (3140)	-55 to 125	8 Ld SOIC Tape and Reel	
CA3140T	-55 to 125	8 Pin Metal Can	T8.C

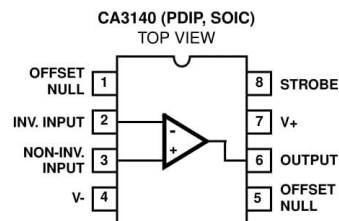
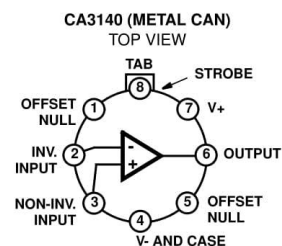
Features

- MOSFET Input Stage
 - Very High Input Impedance (Z_{IN}) - $1.5T\Omega$ (Typ)
 - Very Low Input Current (I_I) - $10pA$ (Typ) at $\pm 15V$
 - Wide Common Mode Input Voltage Range (V_{ICR}) - Can be Swung 0.5V Below Negative Supply Voltage Rail
 - Output Swing Complements Input Common Mode Range
- Directly Replaces Industry Type 741 in Most Applications

Applications

- Ground-Referenced Single Supply Amplifiers in Automobile and Portable Instrumentation
- Sample and Hold Amplifiers
- Long Duration Timers/Multivibrators (μ seconds-Minutes-Hours)
- Photocurrent Instrumentation
- Peak Detectors
- Active Filters
- Comparators
- Interface in 5V TTL Systems and Other Low Supply Voltage Systems
- All Standard Operational Amplifier Applications
- Function Generators
- Tone Controls
- Power Supplies
- Portable Instruments
- Intrusion Alarm Systems

Pinouts



CA3140, CA3140A

Absolute Maximum Ratings

DC Supply Voltage (Between V+ and V- Terminals)	36V
Differential Mode Input Voltage	8V
DC Input Voltage	(V+ +8V) To (V- -0.5V)
Input Terminal Current	1mA
Output Short Circuit Duration (Note 2)	Indefinite

Operating Conditions

Temperature Range	-55°C to 125°C
-------------------	----------------

Thermal Information

Thermal Resistance (Typical, Note 1)	θ_{JA} (°C/W)	θ_{JC} (°C/W)
PDIP Package	100	N/A
SOIC Package	160	N/A
Metal Can Package	170	85
Maximum Junction Temperature (Metal Can Package)	175°C	
Maximum Junction Temperature (Plastic Package)	150°C	
Maximum Storage Temperature Range	-65°C to 150°C	
Maximum Lead Temperature (Soldering 10s)	300°C	
(SOIC - Lead Tips Only)		

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

NOTES:

- θ_{JA} is measured with the component mounted on an evaluation PC board in free air.
- Short circuit may be applied to ground or to either supply.

Electrical Specifications $V_{SUPPLY} = \pm 15V$, $T_A = 25^\circ C$

PARAMETER	SYMBOL	TEST CONDITIONS	TYPICAL VALUES		UNITS
			CA3140	CA3140A	
Input Offset Voltage Adjustment Resistor		Typical Value of Resistor Between Terminals 4 and 5 or 4 and 1 to Adjust Max V_{IO}	4.7	18	k Ω
Input Resistance	R_I		1.5	1.5	T Ω
Input Capacitance	C_I		4	4	pF
Output Resistance	R_O		60	60	Ω
Equivalent Wideband Input Noise Voltage (See Figure 27)	e_N	BW = 140kHz, $R_S = 1M\Omega$	48	48	μV
Equivalent Input Noise Voltage (See Figure 35)	e_N	$R_S = 100\Omega$			
			f = 1kHz	40	nV/ $\sqrt{Hz}$
			f = 10kHz	12	nV/ $\sqrt{Hz}$
Short Circuit Current to Opposite Supply	I_{OM+}	Source	40	40	mA
	I_{OM-}	Sink	18	18	mA
Gain-Bandwidth Product, (See Figures 6, 30)	f_T		4.5	4.5	MHz
Slew Rate, (See Figure 31)	SR		9	9	V/ μs
Sink Current From Terminal 8 To Terminal 4 to Swing Output Low			220	220	μA
Transient Response (See Figure 28)	t_r	$R_L = 2k\Omega$ $C_L = 100pF$	Rise Time	0.08	μs
	OS		Overshoot	10	%
Settling Time at 10V _{P-P} , (See Figure 5)	t_S	$R_L = 2k\Omega$ $C_L = 100pF$ Voltage Follower	To 1mV	4.5	μs
			To 10mV	1.4	μs

Electrical Specifications For Equipment Design, at $V_{SUPPLY} = \pm 15V$, $T_A = 25^\circ C$, Unless Otherwise Specified

PARAMETER	SYMBOL	CA3140			CA3140A			UNITS
		MIN	TYP	MAX	MIN	TYP	MAX	
Input Offset Voltage	$ V_{IO} $	-	5	15	-	2	5	mV
Input Offset Current	$ I_{IO} $	-	0.5	30	-	0.5	20	pA
Input Current	I_I	-	10	50	-	10	40	pA
Large Signal Voltage Gain (Note 3) (See Figures 6, 29)	A_{OL}	20	100	-	20	100	-	kV/V
		86	100	-	86	100	-	dB

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Electrical Specifications For Equipment Design, at $V_{\text{SUPPLY}} = \pm 15\text{V}$, $T_A = 25^\circ\text{C}$, Unless Otherwise Specified (Continued)

PARAMETER	SYMBOL	CA3140			CA3140A			UNITS
		MIN	TYP	MAX	MIN	TYP	MAX	
Common Mode Rejection Ratio (See Figure 34)	CMRR	-	32	320	-	32	320	$\mu\text{V/V}$
		70	90	-	70	90	-	dB
Common Mode Input Voltage Range (See Figure 8)	V_{ICR}	-15	-15.5 to +12.5	11	-15	-15.5 to +12.5	12	V
Power-Supply Rejection Ratio, $\Delta V_{\text{IO}}/\Delta V_{\text{S}}$ (See Figure 36)	PSRR	-	100	150	-	100	150	$\mu\text{V/V}$
		76	80	-	76	80	-	dB
Max Output Voltage (Note 4) (See Figures 2, 8)	$V_{\text{OM}+}$	+12	13	-	+12	13	-	V
	$V_{\text{OM}-}$	-14	-14.4	-	-14	-14.4	-	V
Supply Current (See Figure 32)	I_+	-	4	6	-	4	6	mA
Device Dissipation	P_{D}	-	120	180	-	120	180	mW
Input Offset Voltage Temperature Drift	$\Delta V_{\text{IO}}/\Delta T$	-	8	-	-	6	-	$\mu\text{V}/^\circ\text{C}$

NOTES:

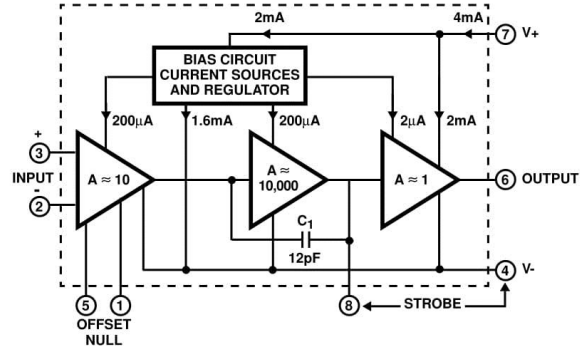
- At $V_{\text{O}} = 26V_{\text{P-P}}$, +12V, -14V and $R_{\text{L}} = 2\text{k}\Omega$.
- At $R_{\text{L}} = 2\text{k}\Omega$.

Electrical Specifications For Design Guidance At $V_+ = 5\text{V}$, $V_- = 0\text{V}$, $T_A = 25^\circ\text{C}$

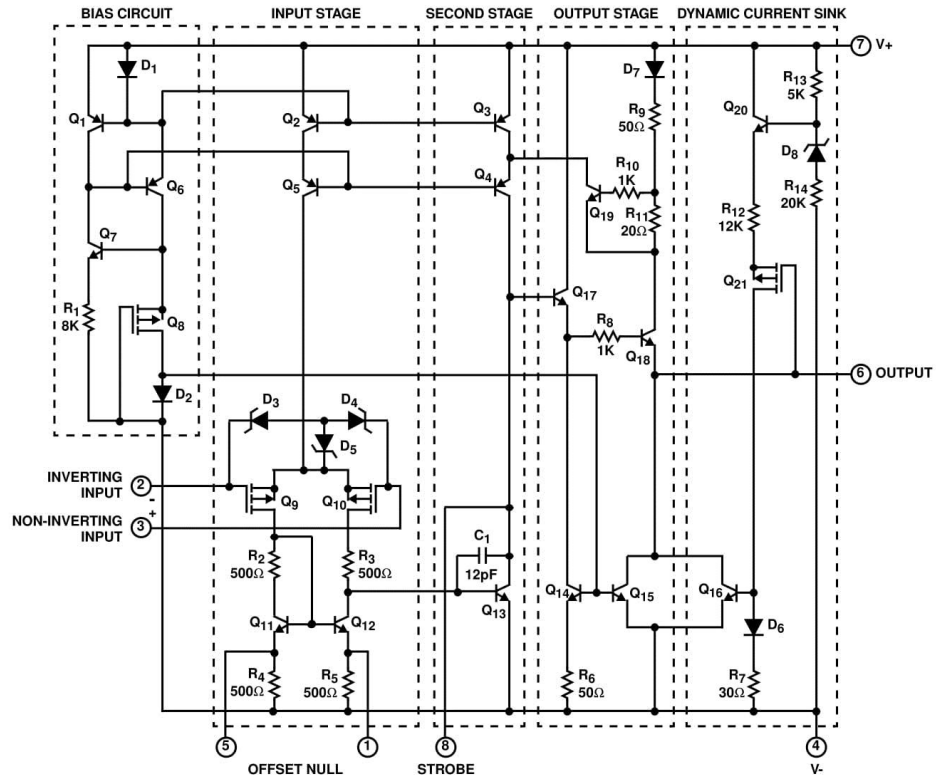
PARAMETER	SYMBOL	TYPICAL VALUES		UNITS
		CA3140	CA3140A	
Input Offset Voltage	$ V_{\text{IO}} $	5	2	mV
Input Offset Current	$ I_{\text{IO}} $	0.1	0.1	pA
Input Current	I_{I}	2	2	pA
Input Resistance	R_{I}	1	1	$\text{T}\Omega$
Large Signal Voltage Gain (See Figures 6, 29)	A_{OL}	100	100	kV/V
		100	100	dB
Common Mode Rejection Ratio	CMRR	32	32	$\mu\text{V/V}$
		90	90	dB
Common Mode Input Voltage Range (See Figure 8)	V_{ICR}	-0.5	-0.5	V
		2.6	2.6	V
Power Supply Rejection Ratio	PSRR $\Delta V_{\text{IO}}/\Delta V_{\text{S}}$	100	100	$\mu\text{V/V}$
		80	80	dB
Maximum Output Voltage (See Figures 2, 8)	$V_{\text{OM}+}$	3	3	V
	$V_{\text{OM}-}$	0.13	0.13	V
Maximum Output Current:	Source $I_{\text{OM}+}$	10	10	mA
	Sink $I_{\text{OM}-}$	1	1	mA
Slew Rate (See Figure 31)	SR	7	7	V/ μs
Gain-Bandwidth Product (See Figure 30)	f_{T}	3.7	3.7	MHz
Supply Current (See Figure 32)	I_+	1.6	1.6	mA
Device Dissipation	P_{D}	8	8	mW
Sink Current from Terminal 8 to Terminal 4 to Swing Output Low		200	200	μA

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Block Diagram



Schematic Diagram



NOTE: All resistance values are in ohms.

Application Information

Circuit Description

As shown in the block diagram, the input terminals may be operated down to 0.5V below the negative supply rail. Two class A amplifier stages provide the voltage gain, and a unique class AB amplifier stage provides the current gain necessary to drive low-impedance loads.

A biasing circuit provides control of cascoded constant current flow circuits in the first and second stages. The CA3140 includes an on chip phase compensating capacitor that is sufficient for the unity gain voltage follower configuration.

Input Stage

The schematic diagram consists of a differential input stage using PMOS field-effect transistors (Q_9 , Q_{10}) working into a mirror pair of bipolar transistors (Q_{11} , Q_{12}) functioning as load resistors together with resistors R_2 through R_5 . The mirror pair transistors also function as a differential-to-single-ended converter to provide base current drive to the second stage bipolar transistor (Q_{13}). Offset nulling, when desired, can be effected with a 10k Ω potentiometer connected across Terminals 1 and 5 and with its slider arm connected to Terminal 4. Cascode-connected bipolar transistors Q_2 , Q_5 are the constant current source for the input stage. The base biasing circuit for the constant current source is described subsequently. The small diodes D_3 , D_4 , D_5 provide gate oxide protection against high voltage transients, e.g., static electricity.

Second Stage

Most of the voltage gain in the CA3140 is provided by the second amplifier stage, consisting of bipolar transistor Q_{13} and its cascode connected load resistance provided by bipolar transistors Q_3 , Q_4 . On-chip phase compensation, sufficient for a majority of the applications is provided by C_1 . Additional Miller-Effect compensation (roll off) can be accomplished, when desired, by simply connecting a small capacitor between Terminals 1 and 8. Terminal 8 is also used to strobe the output stage into quiescence. When terminal 8 is tied to the negative supply rail (Terminal 4) by mechanical or electrical means, the output Terminal 6 swings low, i.e., approximately to Terminal 4 potential.

Output Stage

The CA3140 Series circuits employ a broad band output stage that can sink loads to the negative supply to complement the capability of the PMOS input stage when operating near the negative rail. Quiescent current in the emitter-follower cascade circuit (Q_{17} , Q_{18}) is established by transistors (Q_{14} , Q_{15}) whose base currents are "mirrored" to current flowing through diode D_2 in the bias circuit section. When the CA3140 is operating such that output Terminal 6 is sourcing current, transistor Q_{18} functions as an emitter-follower to source current from the V_+ bus (Terminal 7), via D_7 , R_9 , and R_{11} . Under these conditions, the collector potential of Q_{13} is sufficiently high to permit the necessary flow of base current to emitter follower Q_{17} which, in turn, drives Q_{18} .

When the CA3140 is operating such that output Terminal 6 is sinking current to the V_- bus, transistor Q_{16} is the current sinking element. Transistor Q_{16} is mirror connected to D_6 , R_7 , with current fed by way of Q_{21} , R_{12} , and Q_{20} . Transistor Q_{20} , in turn, is biased by current flow through R_{13} , zener D_8 , and R_{14} . The dynamic current sink is controlled by voltage level sensing. For purposes of explanation, it is assumed that output Terminal 6 is quiescently established at the potential midpoint between the V_+ and V_- supply rails. When output current sinking mode operation is required, the collector potential of transistor Q_{13} is driven below its quiescent level, thereby causing Q_{17} , Q_{18} to decrease the output voltage at Terminal 6. Thus, the gate terminal of PMOS transistor Q_{21} is displaced toward the V_- bus, thereby reducing the channel resistance of Q_{21} . As a consequence, there is an incremental increase in current flow through Q_{20} , R_{12} , Q_{21} , D_6 , R_7 , and the base of Q_{16} . As a result, Q_{16} sinks current from Terminal 6 in direct response to the incremental change in output voltage caused by Q_{18} . This sink current flows regardless of load; any excess current is internally supplied by the emitter-follower Q_{18} . Short circuit protection of the output circuit is provided by Q_{19} , which is driven into conduction by the high voltage drop developed across R_{11} under output short circuit conditions. Under these conditions, the collector of Q_{19} diverts current from Q_4 so as to reduce the base current drive from Q_{17} , thereby limiting current flow in Q_{18} to the short circuited load terminal.

Bias Circuit

Quiescent current in all stages (except the dynamic current sink) of the CA3140 is dependent upon bias current flow in R_1 . The function of the bias circuit is to establish and maintain constant current flow through D_1 , Q_6 , Q_8 and D_2 . D_1 is a diode connected transistor mirror connected in parallel with the base emitter junctions of Q_1 , Q_2 , and Q_3 . D_1 may be considered as a current sampling diode that senses the emitter current of Q_6 and automatically adjusts the base current of Q_6 (via Q_1) to maintain a constant current through Q_6 , Q_8 , D_2 . The base currents in Q_2 , Q_3 are also determined by constant current flow D_1 . Furthermore, current in diode connected transistor Q_2 establishes the currents in transistors Q_{14} and Q_{15} .

Typical Applications

Wide dynamic range of input and output characteristics with the most desirable high input impedance characteristics is achieved in the CA3140 by the use of an unique design based upon the PMOS Bipolar process. Input common mode voltage range and output swing capabilities are complementary, allowing operation with the single supply down to 4V.

The wide dynamic range of these parameters also means that this device is suitable for many single supply applications, such as, for example, where one input is driven below the potential of Terminal 4 and the phase sense of the output signal must be maintained – a most important consideration in comparator applications.

Output Circuit Considerations

Excellent interfacing with TTL circuitry is easily achieved with a single 6.2V zener diode connected to Terminal 8 as shown in Figure 1. This connection assures that the maximum output signal swing will not go more positive than the zener voltage minus two base-to-emitter voltage drops within the CA3140. These voltages are independent of the operating supply voltage.

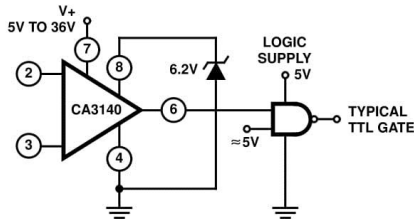


FIGURE 1. ZENER CLAMPING DIODE CONNECTED TO TERMINALS 8 AND 4 TO LIMIT CA3140 OUTPUT SWING TO TTL LEVELS

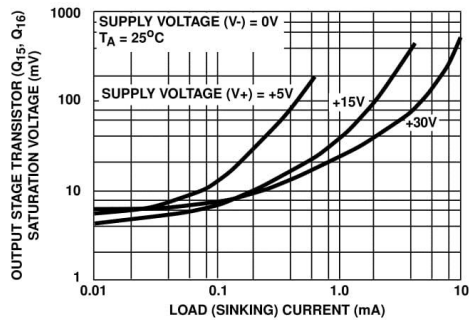


FIGURE 2. VOLTAGE ACROSS OUTPUT TRANSISTORS (Q_{15} AND Q_{16}) vs LOAD CURRENT

Figure 2 shows output current sinking capabilities of the CA3140 at various supply voltages. Output voltage swing to the negative supply rail permits this device to operate both power transistors and thyristors directly without the need for

level shifting circuitry usually associated with the 741 series of operational amplifiers.

Figure 4 shows some typical configurations. Note that a series resistor, R_L , is used in both cases to limit the drive available to the driven device. Moreover, it is recommended that a series diode and shunt diode be used at the thyristor input to prevent large negative transient surges that can appear at the gate of thyristors, from damaging the integrated circuit.

Offset Voltage Nulling

The input offset voltage can be nulled by connecting a $10k\Omega$ potentiometer between Terminals 1 and 5 and returning its wiper arm to terminal 4, see Figure 3A. This technique, however, gives more adjustment range than required and therefore, a considerable portion of the potentiometer rotation is not fully utilized. Typical values of series resistors (R) that may be placed at either end of the potentiometer, see Figure 3B, to optimize its utilization range are given in the Electrical Specifications table.

An alternate system is shown in Figure 3C. This circuit uses only one additional resistor of approximately the value shown in the table. For potentiometers, in which the resistance does not drop to 0Ω at either end of rotation, a value of resistance 10% lower than the values shown in the table should be used.

Low Voltage Operation

Operation at total supply voltages as low as 4V is possible with the CA3140. A current regulator based upon the PMOS threshold voltage maintains reasonable constant operating current and hence consistent performance down to these lower voltages.

The low voltage limitation occurs when the upper extreme of the input common mode voltage range extends down to the voltage at Terminal 4. This limit is reached at a total supply voltage just below 4V. The output voltage range also begins to extend down to the negative supply rail, but is slightly higher than that of the input. Figure 8 shows these characteristics and shows that with 2V dual supplies, the lower extreme of the input common mode voltage range is below ground potential.

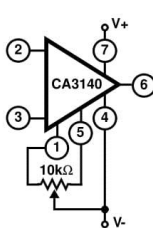


FIGURE 3A. BASIC

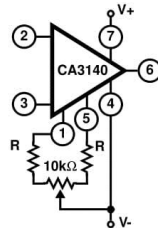


FIGURE 3B. IMPROVED RESOLUTION

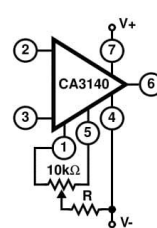


FIGURE 3C. SIMPLER IMPROVED RESOLUTION

FIGURE 3. THREE OFFSET VOLTAGE NULLING METHODS

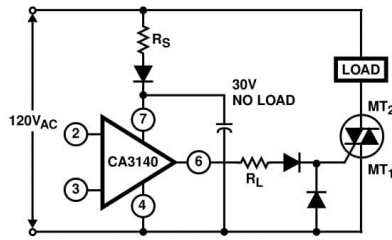
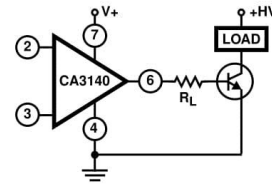

FIGURE 4. METHODS OF UTILIZING THE $V_{CE(SAT)}$ SINKING CURRENT CAPABILITY OF THE CA3140 SERIES


FIGURE 5B. TEST CIRCUITS

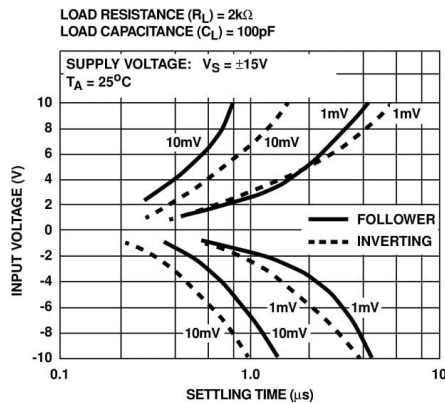


FIGURE 5A. WAVEFORM

FIGURE 5. SETTLING TIME vs INPUT VOLTAGE

Bandwidth and Slew Rate

For those cases where bandwidth reduction is desired, for example, broadband noise reduction, an external capacitor connected between Terminals 1 and 8 can reduce the open loop -3dB bandwidth. The slew rate will, however, also be proportionally reduced by using this additional capacitor. Thus, a 20% reduction in bandwidth by this technique will also reduce the slew rate by about 20%.

Figure 5 shows the typical settling time required to reach 1mV or 10mV of the final value for various levels of large signal inputs for the voltage follower and inverting unity gain amplifiers. The exceptionally fast settling time characteristics

are largely due to the high combination of high gain and wide bandwidth of the CA3140; as shown in Figure 6.

Input Circuit Considerations

As mentioned previously, the amplifier inputs can be driven below the Terminal 4 potential, but a series current limiting resistor is recommended to limit the maximum input terminal current to less than 1mA to prevent damage to the input protection circuitry.

Moreover, some current limiting resistance should be provided between the inverting input and the output when the CA3140 is used as a unity gain voltage follower. This resistance prevents the possibility of extremely large input

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signal transients from forcing a signal through the input protection network and directly driving the internal constant current source which could result in positive feedback via the output terminal. A 3.9k Ω resistor is sufficient.

The typical input current is on the order of 10pA when the inputs are centered at nominal device dissipation. As the output supplies load current, device dissipation will increase, raising the chip temperature and resulting in increased input current. Figure 7 shows typical input terminal current versus ambient temperature for the CA3140.

It is well known that MOSFET devices can exhibit slight changes in characteristics (for example, small changes in input offset voltage) due to the application of large

differential input voltages that are sustained over long periods at elevated temperatures.

Both applied voltage and temperature accelerate these changes. The process is reversible and offset voltage shifts of the opposite polarity reverse the offset. Figure 9 shows the typical offset voltage change as a function of various stress voltages at the maximum rating of 125°C (for metal can); at lower temperatures (metal can and plastic), for example, at 85°C, this change in voltage is considerably less. In typical linear applications, where the differential voltage is small and symmetrical, these incremental changes are of about the same magnitude as those encountered in an operational amplifier employing a bipolar transistor input stage.

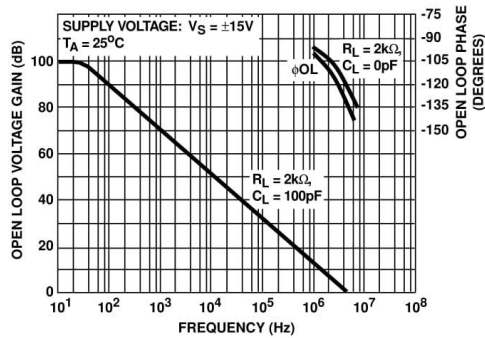


FIGURE 6. OPEN LOOP VOLTAGE GAIN AND PHASE vs FREQUENCY

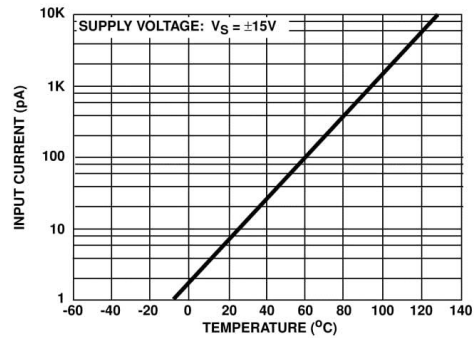


FIGURE 7. INPUT CURRENT vs TEMPERATURE

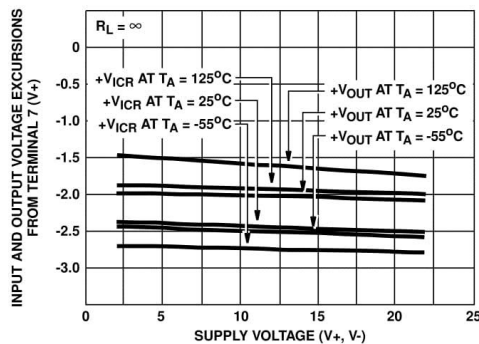
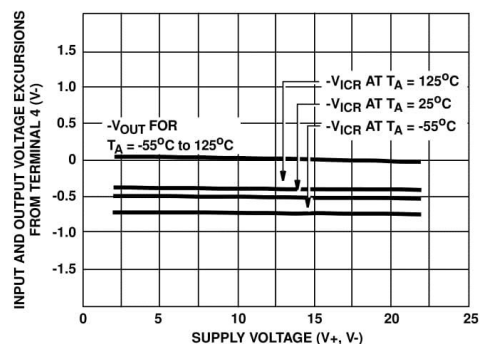


FIGURE 8. OUTPUT VOLTAGE SWING CAPABILITY AND COMMON MODE INPUT VOLTAGE RANGE vs SUPPLY VOLTAGE



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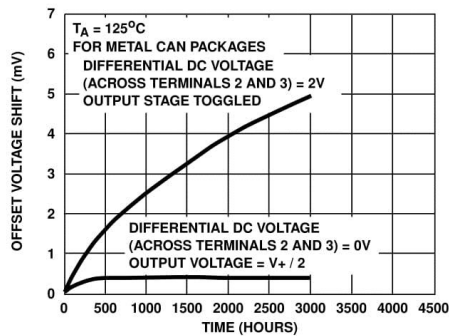


FIGURE 9. TYPICAL INCREMENTAL OFFSET VOLTAGE SHIFT vs OPERATING LIFE

Super Sweep Function Generator

A function generator having a wide tuning range is shown in Figure 10. The 1,000,000/1 adjustment range is accomplished by a single variable potentiometer or by an auxiliary sweeping signal. The CA3140 functions as a non-inverting readout amplifier of the triangular signal developed across the integrating capacitor network connected to the output of the CA3080A current source.

Buffered triangular output signals are then applied to a second CA3080 functioning as a high speed hysteresis switch. Output from the switch is returned directly back to the input of the CA3080A current source, thereby, completing the positive feedback loop.

The triangular output level is determined by the four 1N914 level limiting diodes of the second CA3080 and the resistor divider network connected to Terminal No. 2 (input) of the CA3080. These diodes establish the input trip level to this switching stage and, therefore, indirectly determine the amplitude of the output triangle.

Compensation for propagation delays around the entire loop is provided by one adjustment on the input of the CA3080. This adjustment, which provides for a constant generator amplitude output, is most easily made while the generator is sweeping. High frequency ramp linearity is adjusted by the single 7pF to 60pF capacitor in the output of the CA3080A.

It must be emphasized that only the CA3080A is characterized for maximum output linearity in the current generator function.

Meter Driver and Buffer Amplifier

Figure 11 shows the CA3140 connected as a meter driver and buffer amplifier. Low driving impedance is required of the CA3080A current source to assure smooth operation of the Frequency Adjustment Control. This low-driving impedance requirement is easily met by using a CA3140 connected as a voltage follower. Moreover, a meter may be

placed across the input to the CA3080A to give a logarithmic analog indication of the function generator's frequency.

Analog frequency readout is readily accomplished by the means described above because the output current of the CA3080A varies approximately one decade for each 60mV change in the applied voltage, V_{ABC} (voltage between Terminals 5 and 4 of the CA3080A of the function generator). Therefore, six decades represent 360mV change in V_{ABC} .

Now, only the reference voltage must be established to set the lower limit on the meter. The three remaining transistors from the CA3086 Array used in the sweep generator are used for this reference voltage. In addition, this reference generator arrangement tends to track ambient temperature variations, and thus compensates for the effects of the normal negative temperature coefficient of the CA3080A V_{ABC} terminal voltage.

Another output voltage from the reference generator is used to insure temperature tracking of the lower end of the Frequency Adjustment Potentiometer. A large series resistance simulates a current source, assuring similar temperature coefficients at both ends of the Frequency Adjustment Control.

To calibrate this circuit, set the Frequency Adjustment Potentiometer at its low end. Then adjust the Minimum Frequency Calibration Control for the lowest frequency. To establish the upper frequency limit, set the Frequency Adjustment Potentiometer to its upper end and then adjust the Maximum Frequency Calibration Control for the maximum frequency. Because there is interaction among these controls, repetition of the adjustment procedure may be necessary. Two adjustments are used for the meter. The meter sensitivity control sets the meter scale width of each decade, while the meter position control adjusts the pointer on the scale with negligible effect on the sensitivity adjustment. Thus, the meter sensitivity adjustment control calibrates the meter so that it deflects $1/6$ of full scale for each decade change in frequency.

Sine Wave Shaper

The circuit shown in Figure 12 uses a CA3140 as a voltage follower in combination with diodes from the CA3019 Array to convert the triangular signal from the function generator to a sine-wave output signal having typically less than 2% THD. The basic zero crossing slope is established by the 10k Ω potentiometer connected between Terminals 2 and 6 of the CA3140 and the 9.1k Ω resistor and 10k Ω potentiometer from Terminal 2 to ground. Two break points are established by diodes D_1 through D_4 . Positive feedback via D_5 and D_6 establishes the zero slope at the maximum and minimum levels of the sine wave. This technique is necessary because the voltage follower configuration approaches unity gain rather than the zero gain required to shape the sine wave at the two extremes.

CA3140, CA3140A

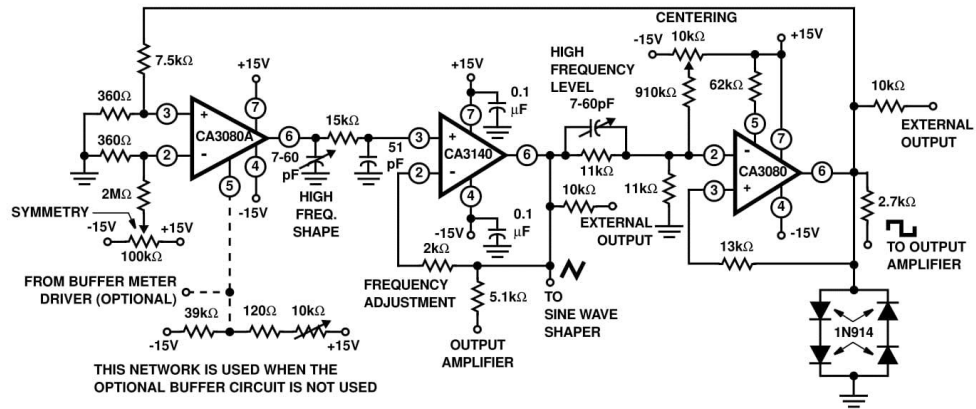
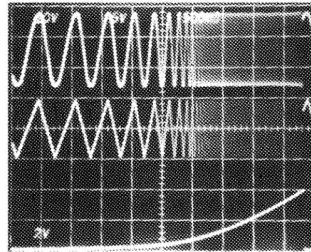
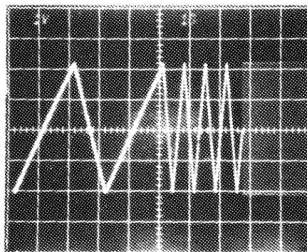


FIGURE 10A. CIRCUIT



Top Trace: Output at junction of 2.7Ω and 51Ω resistors; 5V/Div., 500ms/Div.
Center Trace: External output of triangular function generator; 2V/Div., 500ms/Div.
Bottom Trace: Output of "Log" generator; 10V/Div., 500ms/Div.

FIGURE 10B. FUNCTION GENERATOR SWEEPING



Three tone test signals, highest frequency $\geq 0.5\text{MHz}$. Note the slight asymmetry at the three second/cycle signal. This asymmetry is due to slightly different positive and negative integration from the CA3080A and from the PC board and component leakages at the 100pA level.

FIGURE 10C. FUNCTION GENERATOR WITH FIXED FREQUENCIES

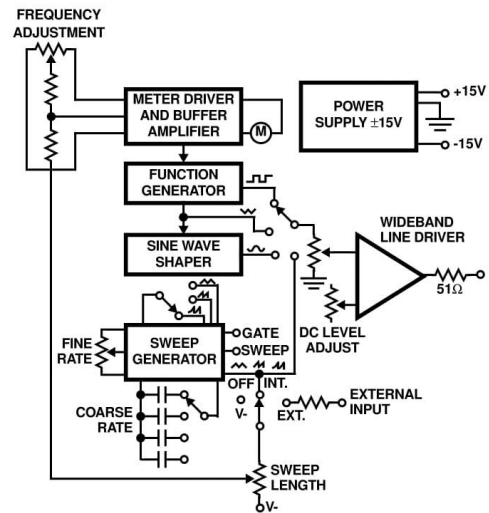


FIGURE 10D. INTERCONNECTIONS

FIGURE 10. FUNCTION GENERATOR

CA3140, CA3140A

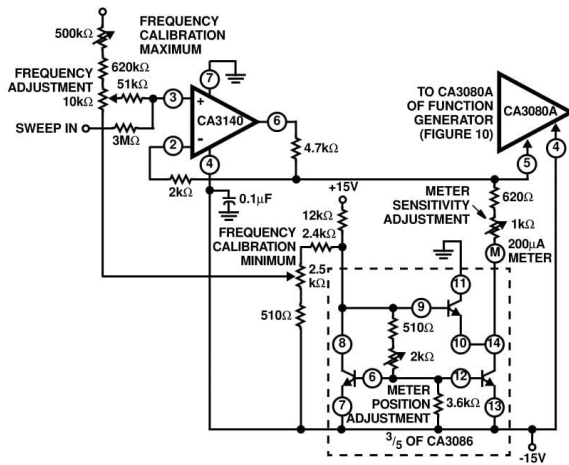


FIGURE 11. METER DRIVER AND BUFFER AMPLIFIER

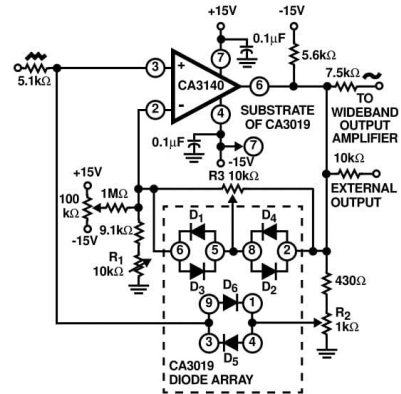


FIGURE 12. SINE WAVE SHAPER

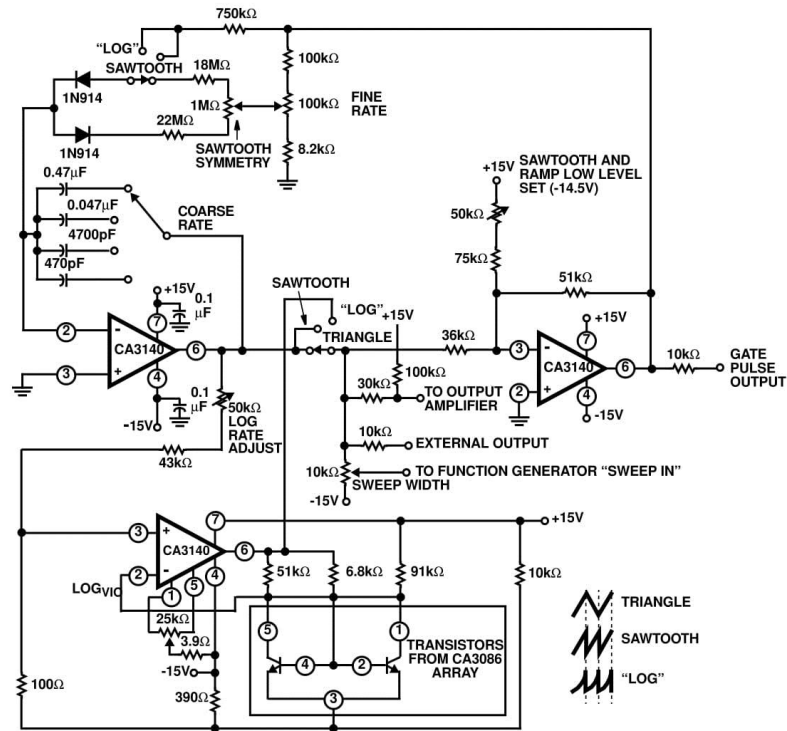


FIGURE 13. SWEEPING GENERATOR

This circuit can be adjusted most easily with a distortion analyzer, but a good first approximation can be made by comparing the output signal with that of a sine wave generator. The initial slope is adjusted with the potentiometer R_1 , followed by an adjustment of R_2 . The final slope is established by adjusting R_3 , thereby adding additional segments that are contributed by these diodes. Because there is some interaction among these controls, repetition of the adjustment procedure may be necessary.

Sweeping Generator

Figure 13 shows a sweeping generator. Three CA3140s are used in this circuit. One CA3140 is used as an integrator, a second device is used as a hysteresis switch that determines the starting and stopping points of the sweep. A third CA3140 is used as a logarithmic shaping network for the log function. Rates and slopes, as well as sawtooth, triangle, and logarithmic sweeps are generated by this circuit.

Wideband Output Amplifier

Figure 14 shows a high slew rate, wideband amplifier suitable for use as a 50Ω transmission line driver. This circuit, when used in conjunction with the function generator and sine wave shaper circuits shown in Figures 10 and 12 provides $18V_{p-p}$ output open circuited, or $9V_{p-p}$ output when terminated in 50Ω . The slew rate required of this amplifier is $28V/\mu s$ ($18V_{p-p} \times \pi \times 0.5MHz$).

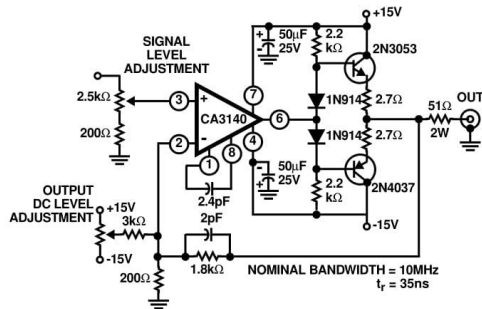


FIGURE 14. WIDEBAND OUTPUT AMPLIFIER

Power Supplies

High input impedance, common mode capability down to the negative supply and high output drive current capability are key factors in the design of wide range output voltage supplies that use a single input voltage to provide a regulated output voltage that can be adjusted from essentially 0V to 24V.

Unlike many regulator systems using comparators having a bipolar transistor input stage, a high impedance reference voltage divider from a single supply can be used in connection with the CA3140 (see Figure 15).

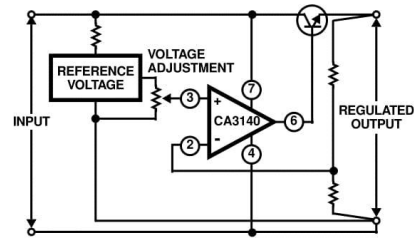


FIGURE 15. BASIC SINGLE SUPPLY VOLTAGE REGULATOR SHOWING VOLTAGE FOLLOWER CONFIGURATION

Essentially, the regulators, shown in Figures 16 and 17, are connected as non inverting power operational amplifiers with a gain of 3.2. An 8V reference input yields a maximum output voltage slightly greater than 25V. As a voltage follower, when the reference input goes to 0V the output will be 0V. Because the offset voltage is also multiplied by the 3.2 gain factor, a potentiometer is needed to null the offset voltage.

Series pass transistors with high I_{CBO} levels will also prevent the output voltage from reaching zero because there is a finite voltage drop (V_{CESAT}) across the output of the CA3140 (see Figure 2). This saturation voltage level may indeed set the lowest voltage obtainable.

The high impedance presented by Terminal 8 is advantageous in effecting current limiting. Thus, only a small signal transistor is required for the current-limit sensing amplifier. Resistive decoupling is provided for this transistor to minimize damage to it or the CA3140 in the event of unusual input or output transients on the supply rail.

Figures 16 and 17, show circuits in which a D2201 high speed diode is used for the current sensor. This diode was chosen for its slightly higher forward voltage drop characteristic, thus giving greater sensitivity. It must be emphasized that heat sinking of this diode is essential to minimize variation of the current trip point due to internal heating of the diode. That is, 1A at 1V forward drop represents one watt which can result in significant regenerative changes in the current trip point as the diode temperature rises. Placing the small signal reference amplifier in the proximity of the current sensing diode also helps minimize the variability in the trip level due to the negative temperature coefficient of the diode. In spite of those limitations, the current limiting point can easily be adjusted over the range from 10mA to 1A with a single adjustment potentiometer. If the temperature stability of the current limiting system is a serious consideration, the more usual current sampling resistor type of circuitry should be employed.

A power Darlington transistor (in a metal can with heatsink), is used as the series pass element for the conventional current limiting system, Figure 16, because high power Darlington dissipation will be encountered at low output voltage and high currents.

CA3140, CA3140A

A small heat sink VERSAWATT transistor is used as the series pass element in the fold back current system, Figure 17, since dissipation levels will only approach 10W. In this system, the D2201 diode is used for current sampling. Foldback is provided by the 3k Ω and 100k Ω divider network connected to the base of the current sensing transistor.

Both regulators provide better than 0.02% load regulation. Because there is constant loop gain at all voltage settings, the

regulation also remains constant. Line regulation is 0.1% per volt. Hum and noise voltage is less than 200 μ V as read with a meter having a 10MHz bandwidth.

Figure 18A shows the turn ON and turn OFF characteristics of both regulators. The slow turn on rise is due to the slow rate of rise of the reference voltage. Figure 18B shows the transient response of the regulator with the switching of a 20 Ω load at 20V output.

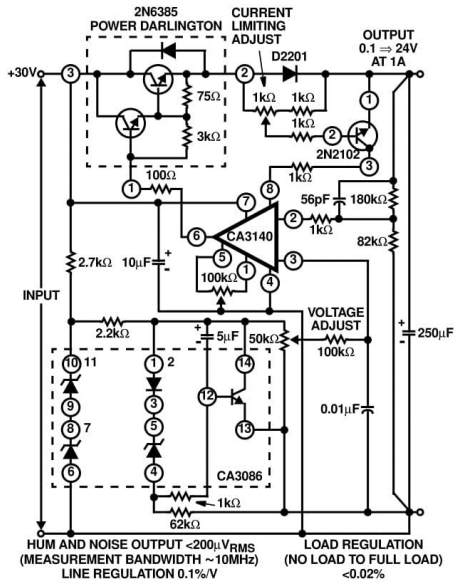


FIGURE 16. REGULATED POWER SUPPLY

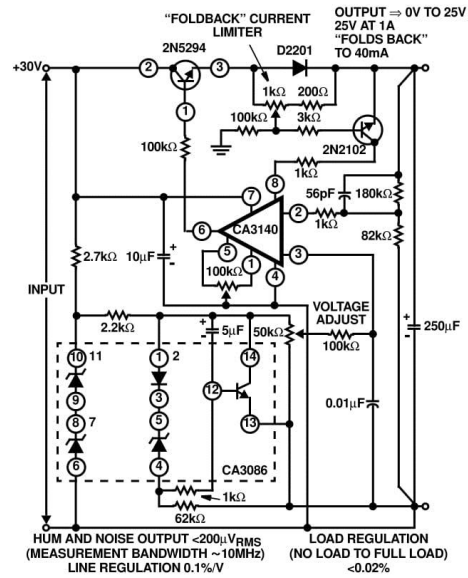
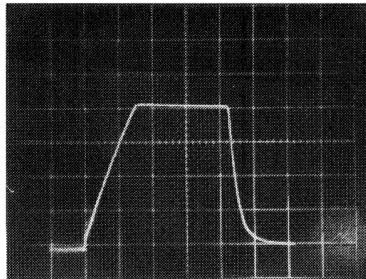
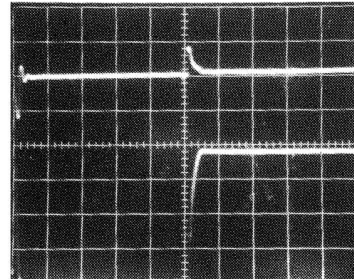


FIGURE 17. REGULATED POWER SUPPLY WITH "FOLDBACK" CURRENT LIMITING



5V/Div., 1s/Div.

FIGURE 18A. SUPPLY TURN-ON AND TUrNOFF CHARACTERISTICS



Top Trace: Output Voltage;
200mV/Div., 5μs/Div.

Bottom Trace: Collector of load switching transistor, load = 1A;
5V/Div., 5μs/Div.

FIGURE 18B. TRANSIENT RESPONSE

FIGURE 18. WAVEFORMS OF DYNAMIC CHARACTERISTICS OF POWER SUPPLY CURRENTS SHOWN IN FIGURES 16 AND 17

CA3140, CA3140A

Tone Control Circuits

High slew rate, wide bandwidth, high output voltage capability and high input impedance are all characteristics required of tone control amplifiers. Two tone control circuits that exploit these characteristics of the CA3140 are shown in Figures 19 and 20.

The first circuit, shown in Figure 20, is the Baxandall tone control circuit which provides unity gain at midband and uses standard linear potentiometers. The high input impedance of the CA3140 makes possible the use of low-cost, low-value, small size capacitors, as well as reduced load of the driving stage.

Bass treble boost and cut are $\pm 15\text{dB}$ at 100Hz and 10kHz , respectively. Full peak-to-peak output is available up to at least 20kHz due to the high slew rate of the CA3140. The amplifier gain is 3dB down from its "flat" position at 70kHz .

Figure 19 shows another tone control circuit with similar boost and cut specifications. The wideband gain of this circuit is equal to the ultimate boost or cut plus one, which in this case is a gain of eleven. For 20dB boost and cut, the input loading of this circuit is essentially equal to the value of the resistance from Terminal No. 3 to ground. A detailed analysis of this circuit is given in "An IC Operational Transconductance Amplifier (OTA) With Power Capability" by L. Kaplan and H. Wittlinger, IEEE Transactions on Broadcast and Television Receivers, Vol. BTR-18, No. 3, August, 1972.

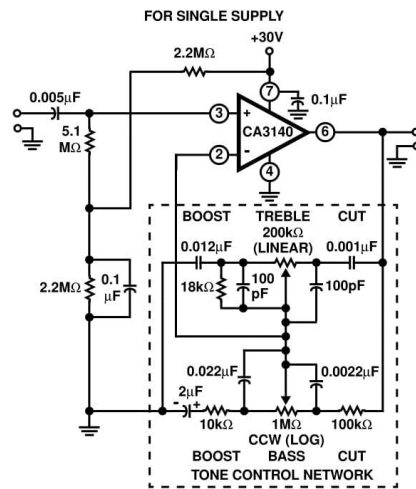
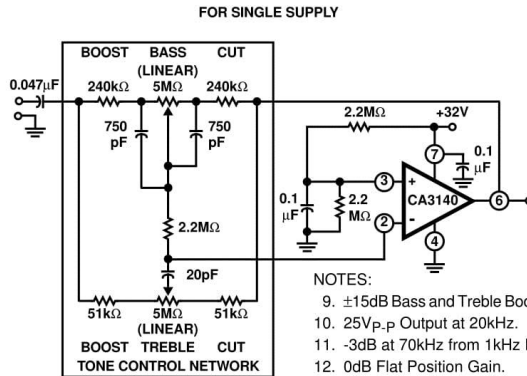
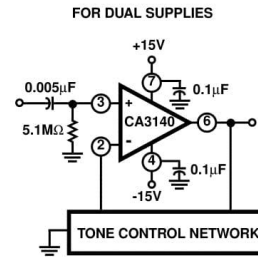


FIGURE 19. TONE CONTROL CIRCUIT USING CA3130 SERIES (20dB MIDBAND GAIN)

NOTES:

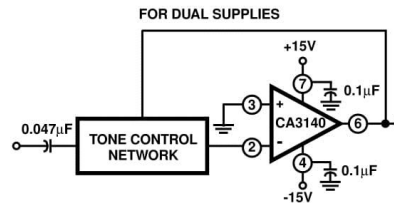
5. 20dB Flat Position Gain.
6. $\pm 15\text{dB}$ Bass and Treble Boost and Cut at 100Hz and 10kHz , respectively.
7. $25\text{V}_{\text{p-p}}$ output at 20kHz .
8. -3dB at 24kHz from 1kHz reference.



NOTES:

9. $\pm 15\text{dB}$ Bass and Treble Boost and Cut at 100Hz and 10kHz , Respectively.
10. $25\text{V}_{\text{p-p}}$ Output at 20kHz .
11. -3dB at 70kHz from 1kHz Reference.
12. 0dB Flat Position Gain.

FIGURE 20. BAXANDALL TONE CONTROL CIRCUIT USING CA3140 SERIES



Wien Bridge Oscillator

Another application of the CA3140 that makes excellent use of its high input impedance, high slew rate, and high voltage qualities is the Wien Bridge sine wave oscillator. A basic Wien Bridge oscillator is shown in Figure 21. When $R_1 = R_2 = R$ and $C_1 = C_2 = C$, the frequency equation reduces to the familiar $f = 1/(2\pi RC)$ and the gain required for oscillation, A_{OSC} is equal to 3. Note that if C_2 is increased by a factor of four and R_2 is reduced by a factor of four, the gain required for oscillation becomes 1.5, thus permitting a potentially higher operating frequency closer to the gain bandwidth product of the CA3140.

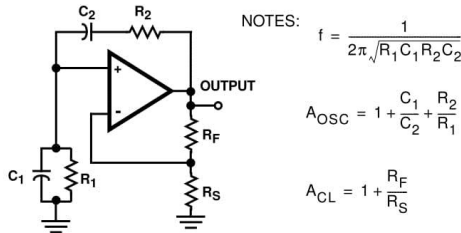


FIGURE 21. BASIC WIEN BRIDGE OSCILLATOR CIRCUIT USING AN OPERATIONAL AMPLIFIER

Oscillator stabilization takes on many forms. It must be precisely set, otherwise the amplitude will either diminish or reach some form of limiting with high levels of distortion. The element, R_S , is commonly replaced with some variable resistance element. Thus, through some control means, the value of R_S is adjusted to maintain constant oscillator output. A FET channel resistance, a thermistor, a lamp bulb, or other device whose resistance increases as the output amplitude is increased are a few of the elements often utilized.

Figure 22 shows another means of stabilizing the oscillator with a zener diode shunting the feedback resistor (R_F of Figure 21). As the output signal amplitude increases, the zener diode impedance decreases resulting in more feedback with consequent reduction in gain; thus stabilizing the amplitude of the output signal. Furthermore, this combination of a monolithic zener diode and bridge rectifier circuit tends to provide a zero temperature coefficient for this regulating system. Because this bridge rectifier system has no time constant, i.e., thermal time constant for the lamp bulb, and RC time constant for filters often used in detector networks, there is no lower frequency limit. For example, with $1\mu F$ polycarbonate capacitors and $22M\Omega$ for the frequency determining network, the operating frequency is 0.007Hz.

As the frequency is increased, the output amplitude must be reduced to prevent the output signal from becoming slew-rate limited. An output frequency of 180kHz will reach a slew rate of approximately 9V/ μs when its amplitude is 16V_{p-p}.

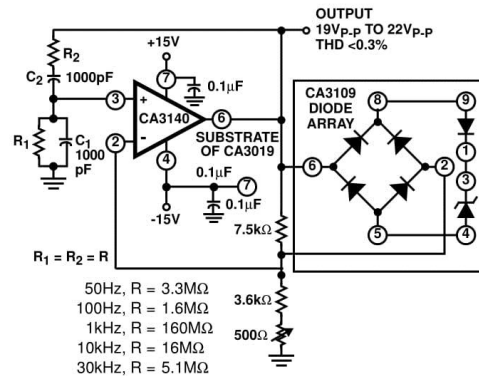


FIGURE 22. WIEN BRIDGE OSCILLATOR CIRCUIT USING CA3140

Simple Sample-and-Hold System

Figure 23 shows a very simple sample-and-hold system using the CA3140 as the readout amplifier for the storage capacitor. The CA3080A serves as both input buffer amplifier and low feed-through transmission switch (see Note 13). System offset nulling is accomplished with the CA3140 via its offset nulling terminals. A typical simulated load of 2kΩ and 30pF is shown in the schematic.

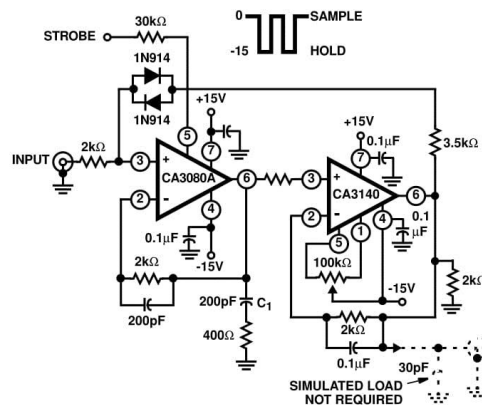


FIGURE 23. SAMPLE AND HOLD CIRCUIT

In this circuit, the storage compensation capacitance (C_1) is only 200pF. Larger value capacitors provide longer "hold" periods but with slower slew rates. The slew rate is:

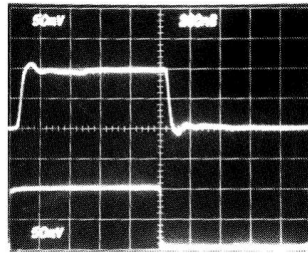
$$\frac{dv}{dt} = \frac{I}{C} = 0.5mA/200pF = 2.5V/\mu s$$

NOTE:

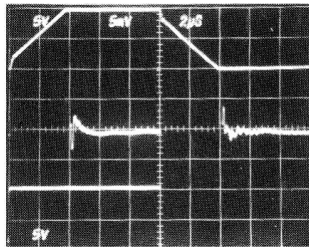
- AN6668 "Applications of the CA3080 and CA 3080A High Performance Operational Transconductance Amplifiers".

CA3140, CA3140A

Pulse "droop" during the hold interval is $170\text{pA}/200\text{pF}$ which is $0.85\mu\text{V}/\mu\text{s}$; (i.e., $170\text{pA}/200\text{pF}$). In this case, 170pA represents the typical leakage current of the CA3080A when strobed off. If C_1 were increased to 2000pF , the "hold-droop" rate will decrease to $0.085\mu\text{V}/\mu\text{s}$, but the slew rate would decrease to $0.25\text{V}/\mu\text{s}$. The parallel diode network connected between Terminal 3 of the CA3080A and Terminal 6 of the CA3140 prevents large input signal feedthrough across the input terminals of the CA3080A to the 200pF storage capacitor when the CA3080A is strobed off. Figure 24 shows dynamic characteristic waveforms of this sample-and-hold system.

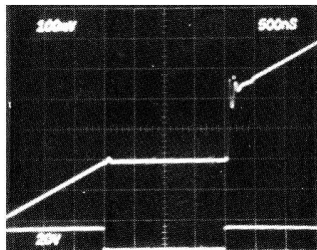


Top Trace: Output; 50mV/Div., 200ns/Div.
Bottom Trace: Input; 50mV/Div., 200ns/Div.



Top Trace: Output Signal; 5V/Div., 2μs/Div.
Center Trace: Difference of Input and Output Signals through Tektronix Amplifier 7A13; 5mV/Div., 2μs/Div.
Bottom Trace: Input Signal; 5V/Div., 2μs/Div.

LARGE SIGNAL RESPONSE AND SETTLING TIME



SAMPLING RESPONSE

Top Trace: Output; 100mV/Div., 500ns/Div.
Bottom Trace: Input; 20V/Div., 500ns/Div.

FIGURE 24. SAMPLE AND HOLD SYSTEM DYNAMIC CHARACTERISTICS WAVEFORMS

Current Amplifier

The low input terminal current needed to drive the CA3140 makes it ideal for use in current amplifier applications such as the one shown in Figure 25 (see Note 14). In this circuit, low current is supplied at the input potential as the power supply to load resistor R_L . This load current is increased by the multiplication factor R_2/R_1 , when the load current is monitored by the power supply meter M. Thus, if the load current is 100nA , with values shown, the load current presented to the supply will be $100\mu\text{A}$; a much easier current to measure in many systems.

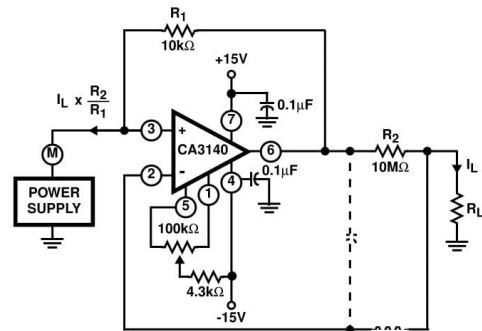


FIGURE 25. BASIC CURRENT AMPLIFIER FOR LOW CURRENT MEASUREMENT SYSTEMS

Note that the input and output voltages are transferred at the same potential and only the output current is multiplied by the scale factor.

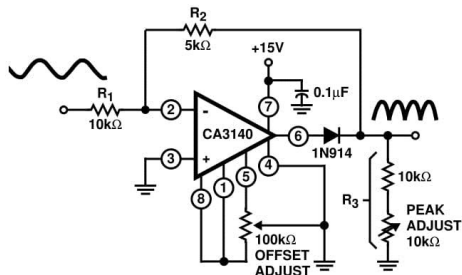
The dotted components show a method of decoupling the circuit from the effects of high output load capacitance and the potential oscillation in this situation. Essentially, the necessary high frequency feedback is provided by the capacitor with the dotted series resistor providing load decoupling.

Full Wave Rectifier

Figure 26 shows a single supply, absolute value, ideal full-wave rectifier with associated waveforms. During positive excursions, the input signal is fed through the feedback network directly to the output. Simultaneously, the positive excursion of the input signal also drives the output terminal (No. 6) of the inverting amplifier in a negative going excursion such that the 1N914 diode effectively disconnects the amplifier from the signal path. During a negative going excursion of the input signal, the CA3140 functions as a normal inverting amplifier with a gain equal to $-R_2/R_1$. When the equality of the two equations shown in Figure 26 is satisfied, the full wave output is symmetrical.

NOTE:

14. "Operational Amplifiers Design and Applications", J. G. Graeme, McGraw-Hill Book Company, page 308, "Negative Immittance Converter Circuits".



$$\text{GAIN} = \frac{R_2}{R_1} = X = \frac{R_3}{R_1 R_2 + R_3}$$

$$R_3 = \left(\frac{X + X^2}{1 - X} \right) R_1$$

$$\text{FOR } X = 0.5 \quad \frac{5\text{k}\Omega}{10\text{k}\Omega} = \frac{R_2}{R_1}$$

$$R_3 = 10\text{k}\Omega \left(\frac{0.75}{0.5} \right) = 15\text{k}\Omega$$

20V_{p-p} Input BW (-3dB) = 290kHz, DC Output (Avg) = 3.2V

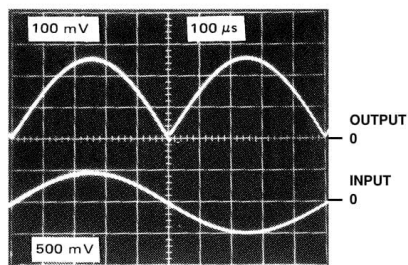
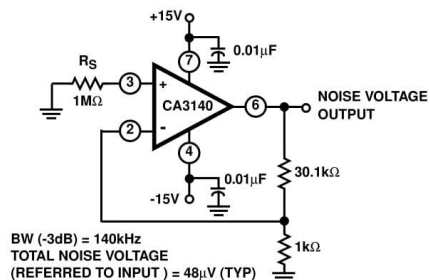


FIGURE 26. SINGLE SUPPLY, ABSOLUTE VALUE, IDEAL FULL WAVE RECTIFIER WITH ASSOCIATED WAVEFORMS



BW (-3dB) = 140kHz
TOTAL NOISE VOLTAGE
(REFERRED TO INPUT) = 48μV (TYP)

FIGURE 27. TEST CIRCUIT AMPLIFIER (30dB GAIN) USED FOR WIDEBAND NOISE MEASUREMENT

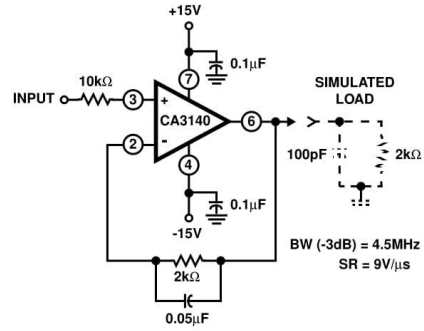
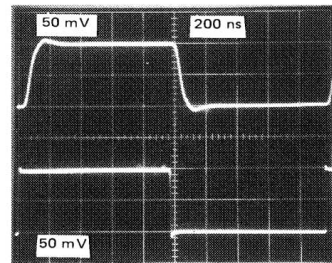
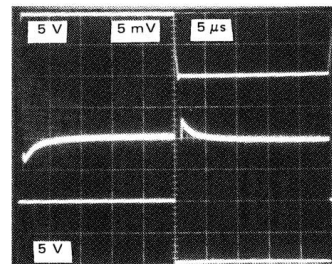


FIGURE 28A. TEST CIRCUIT



Top Trace: Output; 50mV/Div., 200ns/Div.
Bottom Trace: Input; 50mV/Div., 200ns/Div.

FIGURE 28B. SMALL SIGNAL RESPONSE



(Measurement made with Tektronix 7A13 differential amplifier.)

Top Trace: Output Signal; 5V/Div., 5μs/Div.
Center Trace: Difference Signal; 5mV/Div., 5μs/Div.
Bottom Trace: Input Signal; 5V/Div., 5μs/Div.

FIGURE 28C. INPUT-OUTPUT DIFFERENCE SIGNAL SHOWING SETTLING TIME

FIGURE 28. SPLIT SUPPLY VOLTAGE FOLLOWER TEST CIRCUIT AND ASSOCIATED WAVEFORMS

Typical Performance Curves

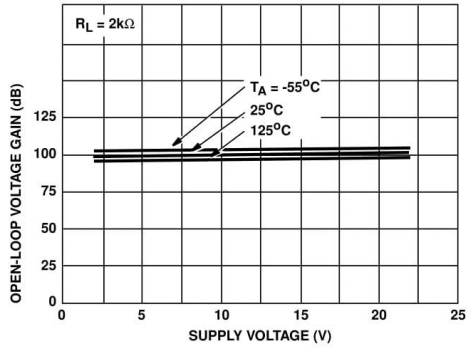


FIGURE 29. OPEN-LOOP VOLTAGE GAIN vs SUPPLY VOLTAGE AND TEMPERATURE

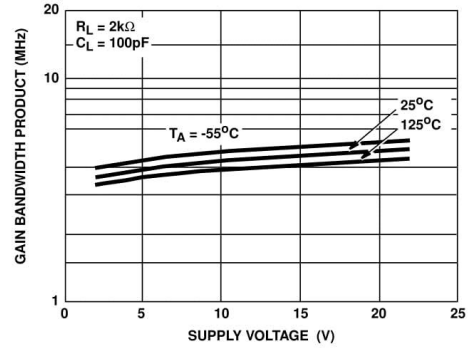


FIGURE 30. GAIN BANDWIDTH PRODUCT vs SUPPLY VOLTAGE AND TEMPERATURE

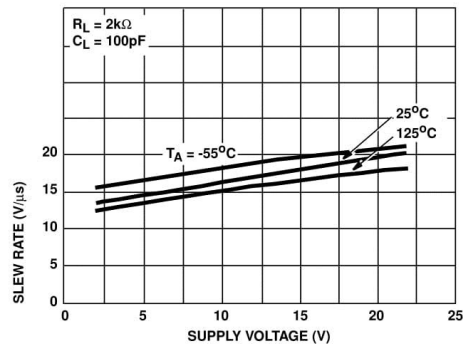


FIGURE 31. SLEW RATE vs SUPPLY VOLTAGE AND TEMPERATURE

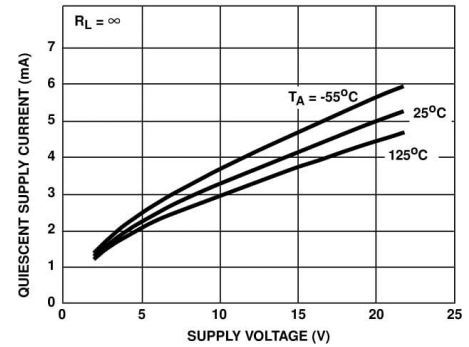


FIGURE 32. QUIESCENT SUPPLY CURRENT vs SUPPLY VOLTAGE AND TEMPERATURE

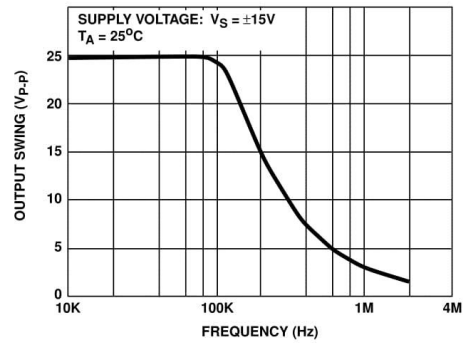


FIGURE 33. MAXIMUM OUTPUT VOLTAGE SWING vs FREQUENCY

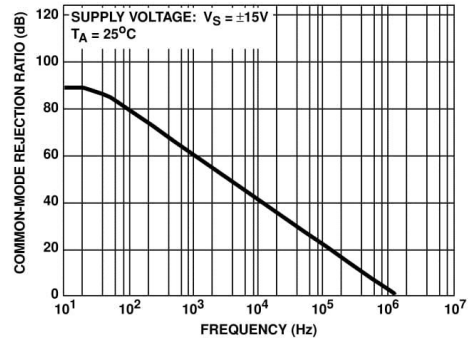


FIGURE 34. COMMON MODE REJECTION RATIO vs FREQUENCY

Typical Performance Curves (Continued)

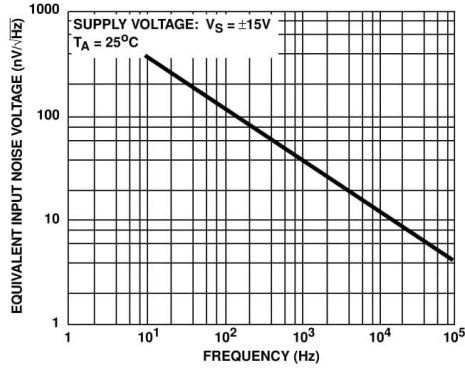


FIGURE 35. EQUIVALENT INPUT NOISE VOLTAGE vs FREQUENCY

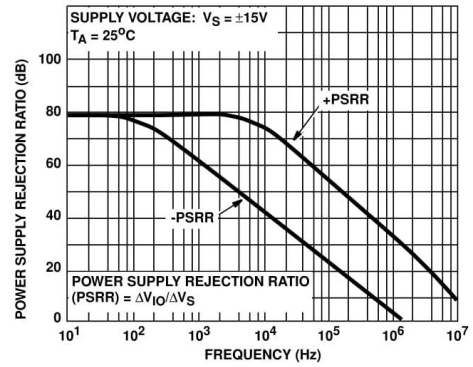
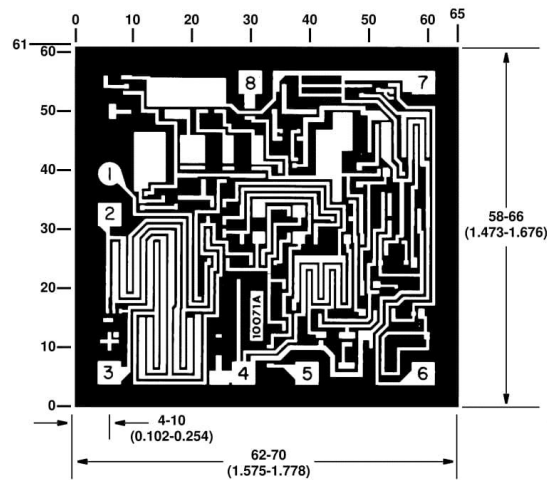


FIGURE 36. POWER SUPPLY REJECTION RATIO vs FREQUENCY

Metallization Mask Layout



Dimensions in parenthesis are in millimeters and are derived from the basic inch dimensions as indicated. Grid graduations are in mils (10^{-3} inch).

The photographs and dimensions represent a chip when it is part of the wafer. When the wafer is cut into chips, the cleavage angles are 57° instead of 90° with respect to the face of the chip. Therefore, the isolated chip is actually 7 mils (0.17mm) larger in both dimensions.