

AMPLIFICATEUR DIFFÉRENTIEL DE TRANSCONDUCTANCE LM13600 ¹

PRESENTATION DU CIRCUIT

On appelle amplificateur de transconductance (ou convertisseur tension courant), un circuit actif dont le courant de sortie est proportionnel à la tension d'entrée (le facteur de proportionnalité est la transconductance g_m).

Le circuit intégré LM13600 comporte deux amplificateurs différentiels de transconductances ajustables électriquement par l'intermédiaire d'un courant de référence I_{ref} . La figure 1 représente un de ces amplificateurs (TR_1 ou TR_2).

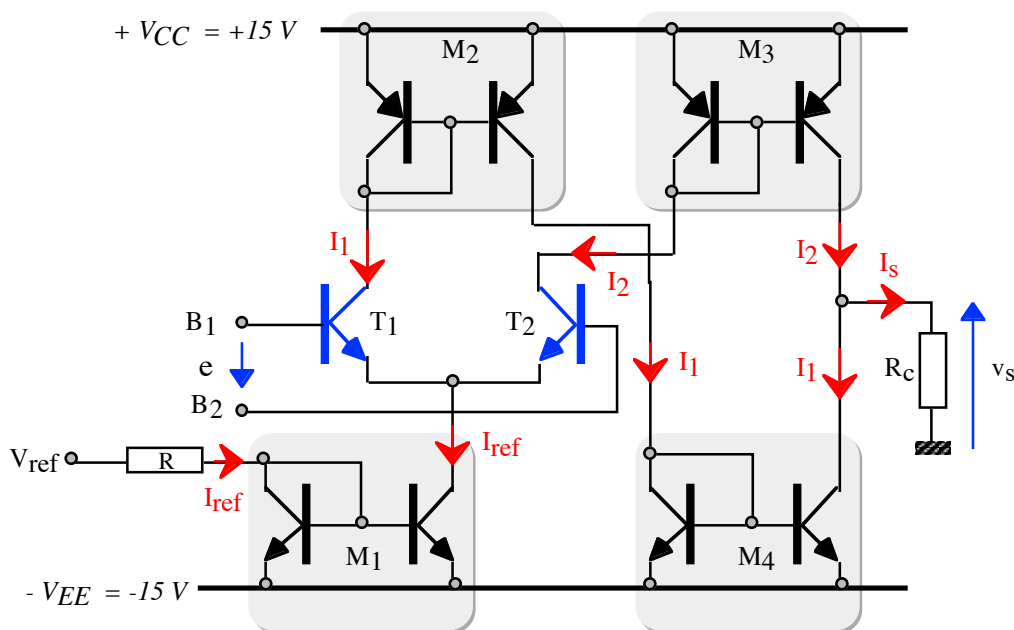


Figure 1 : Schéma simplifié de TR_1 (ou TR_2).

Les transistors identiques T_1 et T_2 de la figure 1 forment une paire différentielle. Leur courant de collecteur (respectivement I_1 et I_2), sont les courants de référence des miroirs de courant M_2 et M_3 qui utilisent des transistors identiques PNP.

Le miroir de courant NPN M_4 associé à M_2 , assure la recopie du courant I_1 . Le courant de sortie I_s est alors : $I_s = I_2 - I_1$.

Le courant de référence I_{ref} fourni par la résistance externe R extérieure au circuit intégré et la tension d'alimentation continue V_{ref} , pourvoit à la transconductance variable $g_m = f(I_{ref})$ du montage.

Au repos, pour une tension différentielle d'entrée e nulle appliquée entre B_1 et B_2 , les courants I_1 et I_2 sont égaux à $I_{ref}/2$ car les tensions V_{BE1} et V_{BE2} de T_1 et T_2 sont égales.

¹ Ph.ROUX © 2009

Lorsqu'on applique une tension continue e positive entre les bases B₂ et B₁ de T₂ et T₁, la relation : $e = V_{BE2} - V_{BE1} > 0$ entraîne $V_{BE2} > V_{BE1}$. Dans ces conditions les courants continus I₁ et I₂ sont tels que :

$$I_2 = \frac{I_{ref}}{2} + \Delta I \quad (1) \quad I_1 = \frac{I_{ref}}{2} - \Delta I \quad (2)$$

Où ΔI représente une petite variation de courant.

Sachant que : $I_1 = I_{SBC} \exp(\frac{V_{BE1}}{U_T})$ et $I_2 = I_{SBC} \exp(\frac{V_{BE2}}{U_T})$

On en déduit la relation : $\frac{I_1}{I_2} = \exp(\frac{e}{U_T})$ qui donne : $e = U_T \ln(\frac{I_2}{I_1})$

Compte tenu des relations (1) et (2), on obtient : $\frac{e}{U_T} = \ln[1 + \frac{2\Delta I}{I_{ref}}] - \ln[1 - \frac{2\Delta I}{I_{ref}}]$

En développant la fonction ln(1 ± ε) au 2^o ordre soit : ± ε - ε²/2, il vient alors : $I_S = I_2 - I_1 = \frac{I_{ref}}{2U_T} e$

Le courant de sortie I_S du montage est proportionnel à la tension différentielle d'entrée e. La transconductance g_m(I_{ref}) est telle que :

$$g_m(I_{ref}) = \frac{I_{ref}}{2U_T}$$

La transconductance g_m dépendant de $U_T = \frac{kT}{q}$ soit 25 mV à 25°C est sensible à la température.

Pour minimiser ce phénomène, il est alors nécessaire de construire une source de courant I_{ref} ayant la même dérive en température.

Exemple d'application : Amplificateur à gain ajustable électriquement.

Considérons le montage donné en figure 2 à 25°C.

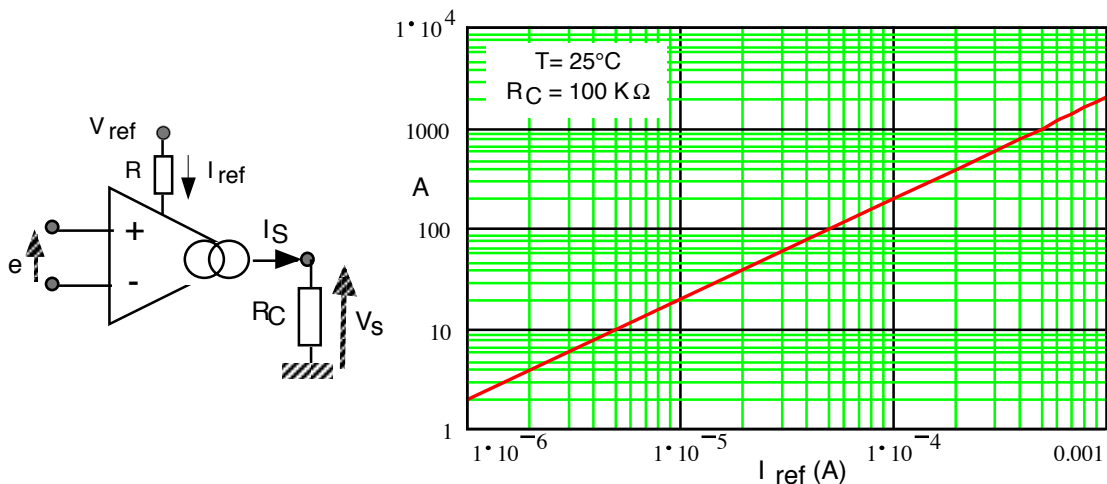


Figure 2 : Amplificateur de tension à gain commandé par I_{ref}.

Le gain en tension du montage tel que : $A(I_{ref}) = \frac{v_s}{e} = \frac{R_C}{2U_T} I_{ref}$

CIRCUIT ANNEXE DU CIRCUIT INTEGRE LM 13600

Le circuit intégré LM13600 possède deux circuits supplémentaires du type « suiveur de tension » donnés en figure 3. Il s'agit d'un montage du type collecteur commun utilisant deux transistors montés en « Darlington ». Ce montage possède les propriétés suivantes :

- Courant d'entrée négligeable
- Résistance d'entrée très grande.
- Gain en tension égale à un

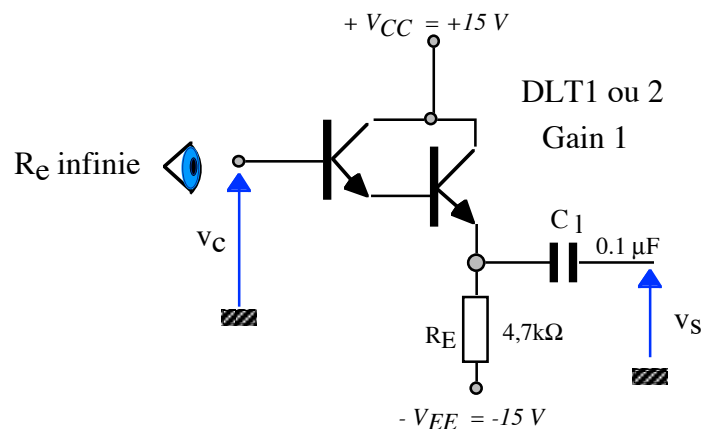


Figure 3 : Montage annexe de type « suiveur »

EXERCICE 1 : FILTRE PASSE-BAS DU 1° ORDRE AVEC ACCORD ELECTRIQUE

Le montage de la figure 4 représente un filtre passe-bas du premier ordre . On admettra que le courant des entrées + et - de TR1 est négligeable et que la résistance d'entrée de l'amplificateur de transconductance TR1 est infinie. La tension de référence V_{ref} est nulle et la température est fixée à $T = 25^\circ\text{C}$.

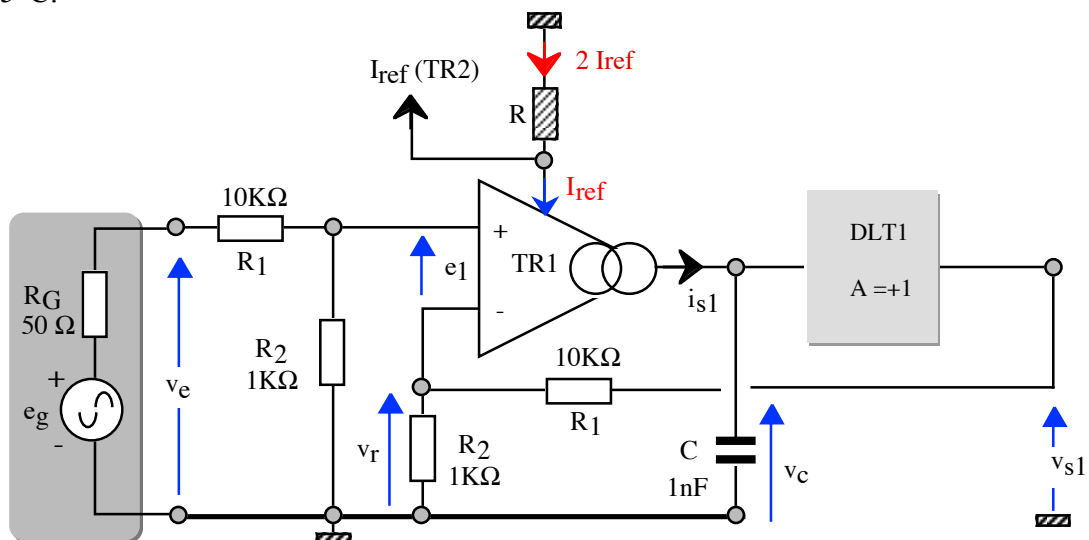


Figure 4 : filtre passe-bas du premier ordre.

1. Compte tenu des propriétés des amplificateurs TR1 et DLT1 développées en première partie, montrer que le gain en tension du montage peut se mettre sous la forme :

$$\frac{v_{s1}}{v_e} = \frac{1}{1 + j \frac{f}{f_c(I_{ref})}}$$

On posera pour effectuer le calcul : $a = \frac{R_2}{R_1 + R_2}$

2. Donner l'expression de la fréquence de coupure f_c du filtre en fonction de a , C , U_T et I_{ref} .
3. Tracer le graphe donnant la fréquence de coupure haute f_h en fonction du courant I_{ref} en coordonnées log-log.
4. On fixe I_{ref} à $10 \mu A$.
 - a. Calculer la valeur de la résistance R qui fixe I_{ref} .
 - b. Tracer sur papier semi-logarithmique, le graphe asymptotique de Bode du module de la fonction de transfert du filtre en fonction de la fréquence réduite : $x = \frac{f}{f_c(I_{ref})}$.

EXERCICE 2 : FILTRE PASSE-BANDE A ACCORD ELECTRIQUE

On obtient un filtre passe-bande accordable électriquement en utilisant deux filtres du 1^o ordre bouclés comme indiqué en figure 2.

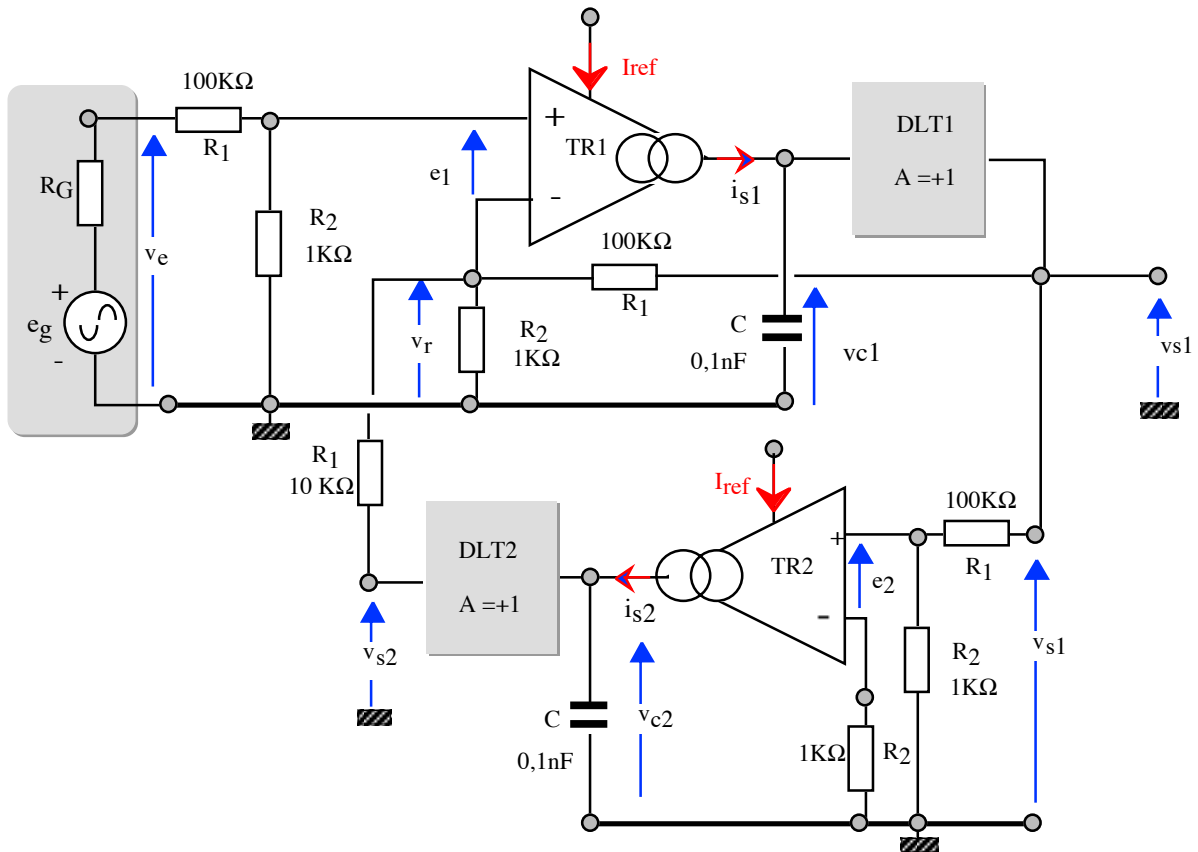


Figure 2 : Filtre passe-bande

1. Pour déterminer l'expression de la fonction de transfert du filtre, on recherchera les relations liant les tensions :
 - a. e_1 et v_{S1} en fonction de g_m , C et ω .
 - b. v_{S2} et v_{S1} en fonction de R_1 , R_2 , g_m , C et ω .
 - c. e_1 , v_e et v_r en fonction de R_1 et R_2 .
 - d. v_r en fonction de v_{S1} et v_{S2} en utilisant le théorème de superposition et en assimilant en première approximation : $[R_1/R_2]$ à R_2 sachant que : $R_1 \gg R_2$.

Montrer que la fonction de transfert est telle que :

$$\frac{v_{s1}}{v_e} = \frac{I}{1 + j\omega \frac{C(R_1 + R_2)}{g_m R_2} + \frac{I}{j\omega C(R_1 + R_2)} gmR_2} \quad (1)$$

2. Déterminer la fréquence de résonance $f_0(I_{ref})$ pour laquelle la partie imaginaire de la relation (1) s'annule. Tracer le graphe de $f_0(I_{ref})$ pour : $0,1\mu A < I_{ref} < 100\mu A$.
3. Montrer que la fonction de transfert peut se mettre sous la forme :

$$\frac{v_{s1}}{v_e} = \frac{I}{1 + j\left(\frac{f}{f_0(I_{ref})} - \frac{f_0(I_{ref})}{f}\right)}$$

4. On fixe I_{ref} à $50\mu A$. Tracer sur papier semi-logarithmique, le graphe asymptotique de Bode du module de la fonction de transfert du filtre en fonction de la fréquence réduite :

$$x = \frac{f}{f_0(I_{ref})}$$

EXERCICE 3 : SORTIE v_{S2} DU FILTRE PASSE-BANDE

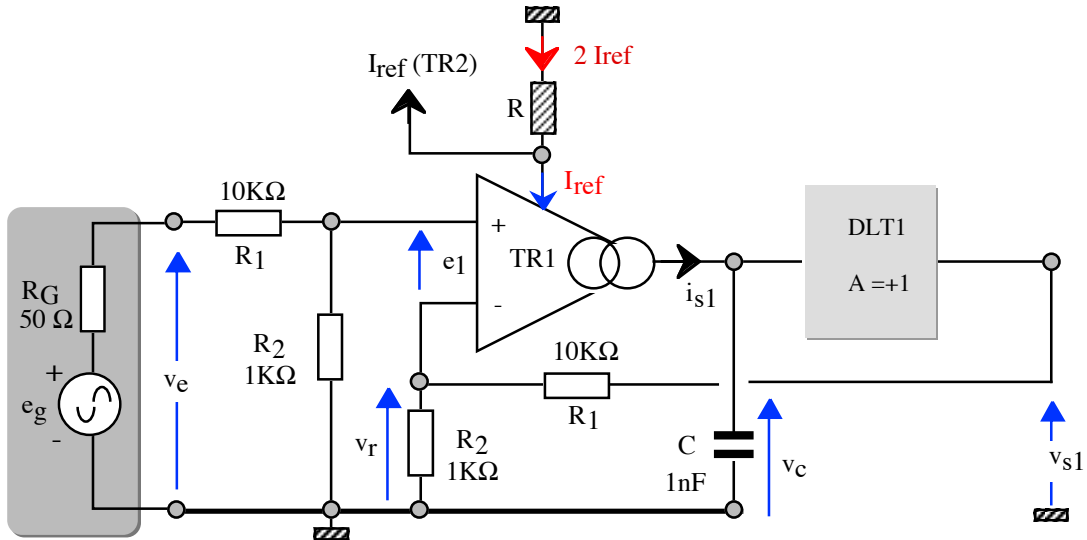
1. On s'intéresse à la sortie v_{S2} du montage de la figure 2. En exploitant les résultats de l'exercice précédent, montrer que la fonction de transfert du filtre en fréquence réduite, est telle que :

$$\frac{v_{s2}}{v_e} = \frac{\frac{I}{jx}}{1 + j\left(x - \frac{I}{x}\right)}$$

2. Tracer sur papier semi-logarithmique, le graphe asymptotique de Bode du module de la fonction de transfert du filtre en fonction de la fréquence réduite. Conclure.

CORRECTION

EXERCICE 1 : FILTRE PASSE-BAS DU 1° ORDRE AVEC ACCORD ELECTRIQUE



1. Gain en tension du montage.

$$e_1 = v_e \frac{R_2}{R_1 + R_2} - v_r \quad \text{soit : } e_1 = av_e - v_r \quad \text{avec : } v_r = av_{s1}$$

$$v_c = \frac{i_{s1}}{j\omega C} = \frac{g_m(I_{ref})}{j\omega C} e_1 \quad \text{avec : } v_{s1} = v_c$$

Il vient alors :

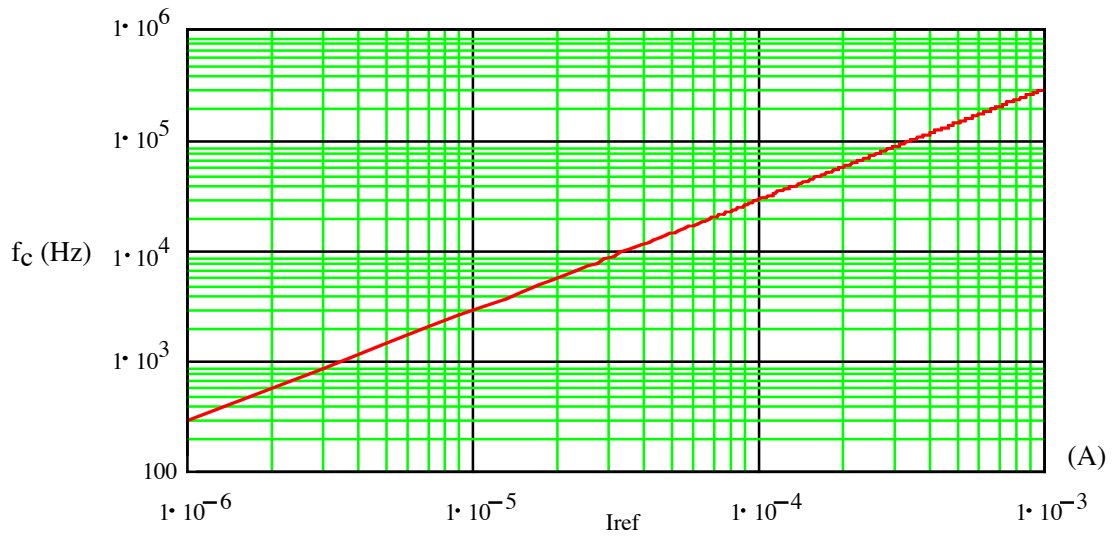
$$v_{s1} = \frac{g_m(I_{ref})}{j\omega C} av_e - \frac{g_m(I_{ref})}{j\omega C} v_{s1}$$

$$\frac{v_{s1}}{v_e} = \frac{I}{1 + j\omega \frac{C}{ag_m(I_{ref})}}$$

2. L'expression du gain montre que le montage se comporte comme un filtre passe-bas du premier ordre ayant pour fonction de transfert : $\frac{v_{s1}}{v_e} = \frac{I}{1 + j\frac{\omega}{\omega_c}}$

La fréquence de coupure à -3 dB est alors : $f_c(I_{ref}) = \frac{1}{2\pi} \frac{ag_m(I_{ref})}{C}$

3. Graphe de f_c en fonction de I_{ref} :



4.

a) Valeur de la résistance $R = 720 \text{ K}\Omega$.

b) En fonction de la fréquence réduite, la fonction de transfert du filtre s'exprime :

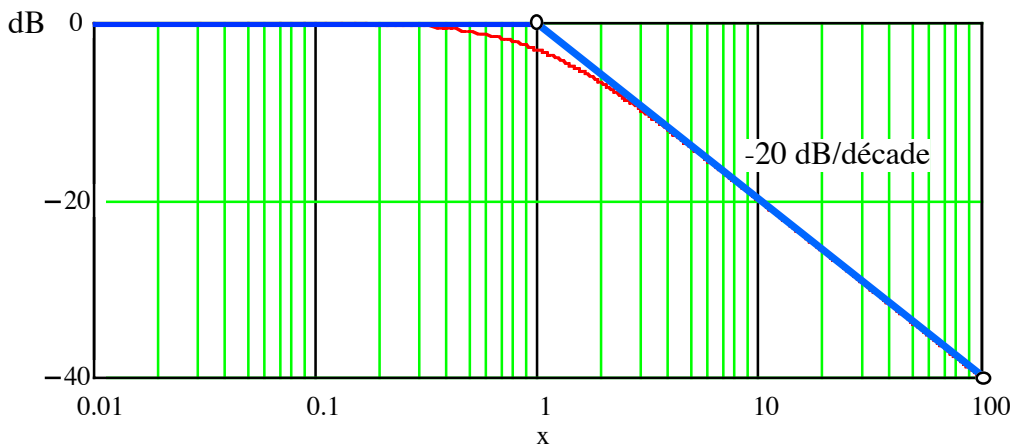
$$\frac{v_{sl}}{v_e} = \frac{I}{I + jx} \quad \text{soit :} \quad \left| \frac{v_{sl}}{v_e} \right|_{dB} = -10 \log(1 + x^2)$$

$$\text{Pour } x = 1 \quad \left| \frac{v_{sl}}{v_e} \right|_{dB} = -3 \text{ dB}$$

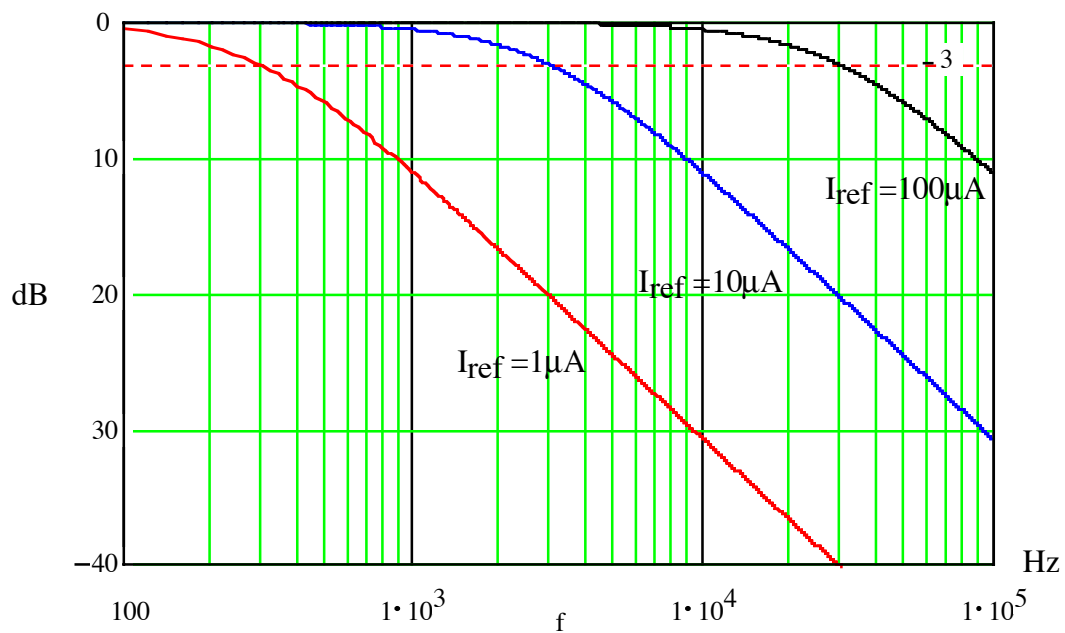
$$\text{Pour } x \rightarrow 0 \quad \left| \frac{v_{sl}}{v_e} \right|_{dB} \rightarrow 0 \text{ dB}$$

$$\text{Pour } x \rightarrow \infty \quad \left| \frac{v_{sl}}{v_e} \right|_{dB} \approx -20 \log x \quad \text{le graphe passe par } x = 1 \text{ avec un coefficient directeur de } -20 \text{ dB/décade.}$$

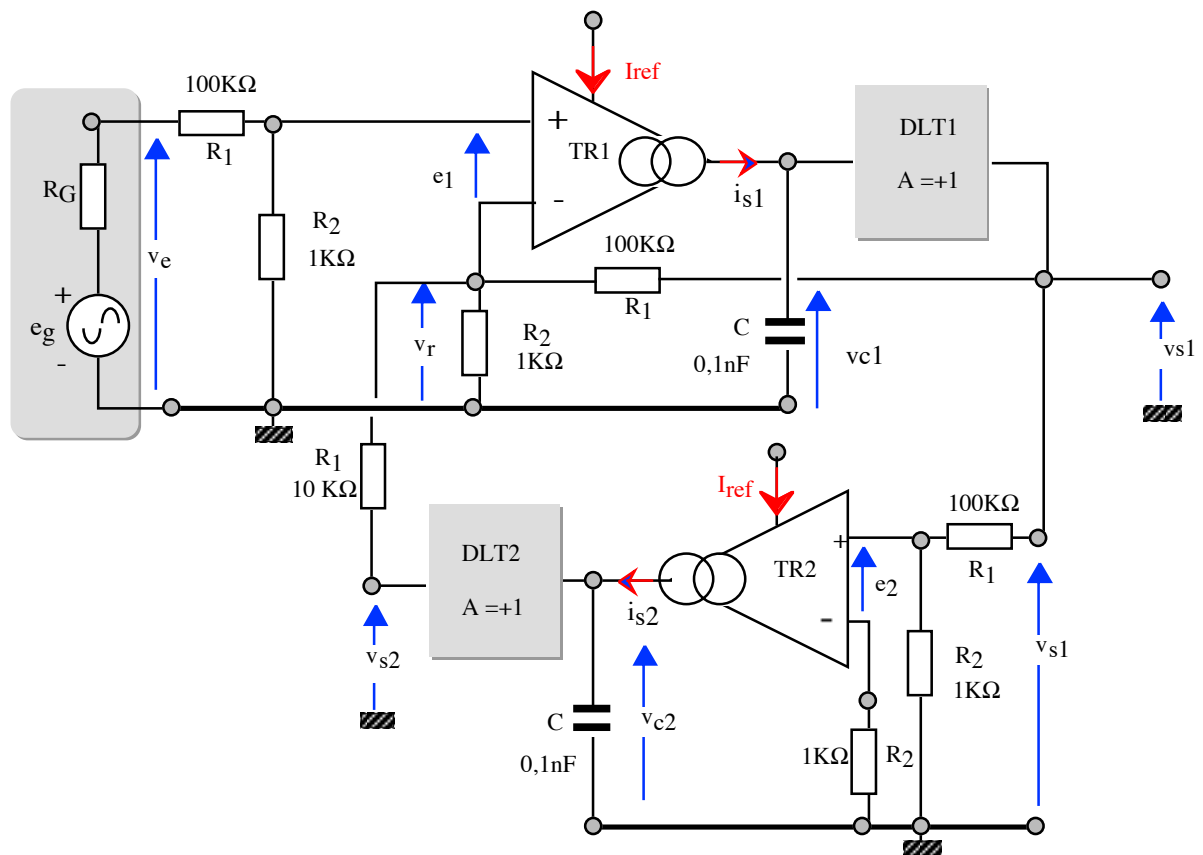
Graphe asymptotique de Bode du module de la fonction de transfert du filtre



Courbes de réponse en fréquences du filtre passe-bas pour 3 valeurs du courant I_{ref} .



2° PARTIE : FILTRE PASSE-BANDE A ACCORD ELECTRIQUE



1. Fonction de transfert du filtre.
 - a. e_1 et v_{s1} en fonction de g_m , C et ω .

$$v_{c1} = v_{s1} = e_1 i_{s1} = e_1 \frac{g_m}{j\omega C} \quad (1)$$

b. Sachant que le courant sur l'entrée – de TR2 est nul :

$$v_{s2} = v_{c2} = \frac{g_m}{j\omega C} v_{s1} \frac{R_2}{R_1 + R_2} \quad (2)$$

c. e_1 , v_e et v_r en fonction de R_1 et R_2 .

$$e_1 = v_e \frac{R_2}{R_1 + R_2} - v_r \quad (3)$$

d. En utilisant le théorème de superposition et en assimilant en première approximation : $[R_1/R_2]$ à R_2 .

$$v_r = \frac{R_2}{R_1 + R_2} v_{s2} + \frac{R_2}{R_1 + R_2} v_{s1} \quad (4)$$

L'équation (1) devient alors :

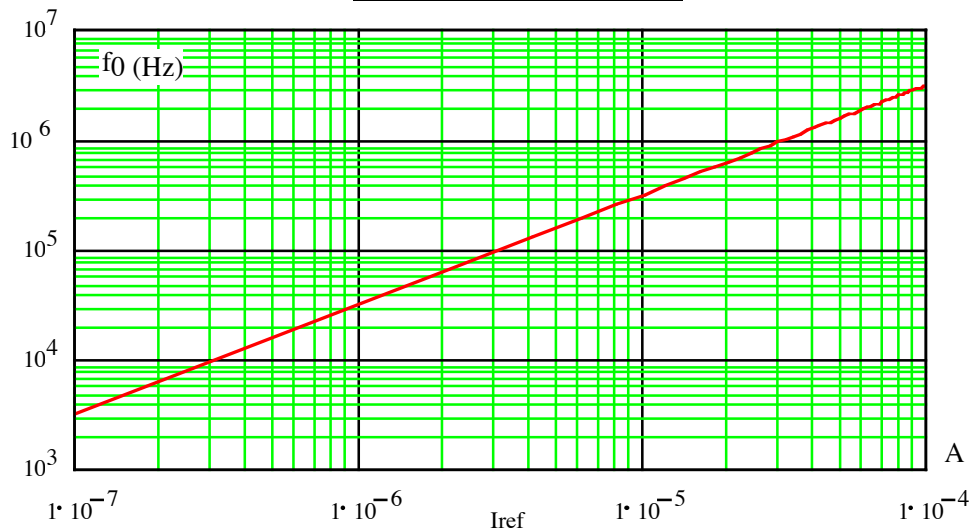
$$v_{s1} = \frac{g_m}{j\omega C} \left(\frac{R_2}{R_1 + R_2} v_e - \frac{R_2^2}{(R_1 + R_2)^2} \frac{g_m}{j\omega C} v_{s1} \frac{R_2}{R_1 + R_2} - \frac{R_2}{R_1 + R_2} v_{s1} \right)$$

Après avoir effectuer des regroupements, on obtient finalement l'expression :

$$\frac{v_{s1}}{v_e} = \frac{I}{1 + j\omega \frac{C(R_1 + R_2)}{g_m R_2} + \frac{I}{j\omega C(R_1 + R_2)} \frac{g_m R_2}{g_m R_2}}$$

2. Fréquence de résonance f_0 (I_{ref}) pour laquelle la partie imaginaire de la relation (1) est nulle :

$$f_0(I_{ref}) = \frac{1}{2\pi} \frac{R_2 g_m(I_{ref})}{(R_1 + R_2) C}$$



Graph de f_0 en fonction de I_{ref}

3. Il est facile de mettre la relation (1) sous la forme :

$$\frac{v_{s1}}{v_e} = \frac{I}{1 + j \left(\frac{f}{f_0(I_{ref})} - \frac{f_0(I_{ref})}{f} \right)}$$

4. Expression de la fonction de transfert :

$$\left| \frac{v_{sl}}{v_e} \right|_{dB} = -10 \log \left(1 + \left(x - \frac{1}{x} \right)^2 \right)$$

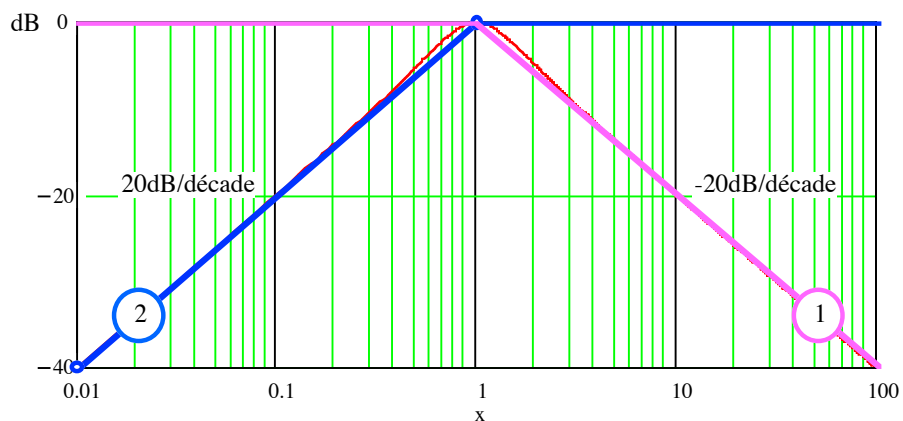
Pour $x = 1$ $\left| \frac{v_{sl}}{v_e} \right|_{dB} = -3dB$

Pour $x \gg 1$ on obtient : $\left| \frac{v_{sl}}{v_e} \right|_{dB} \approx -10 \log(1 + (x)^2)$

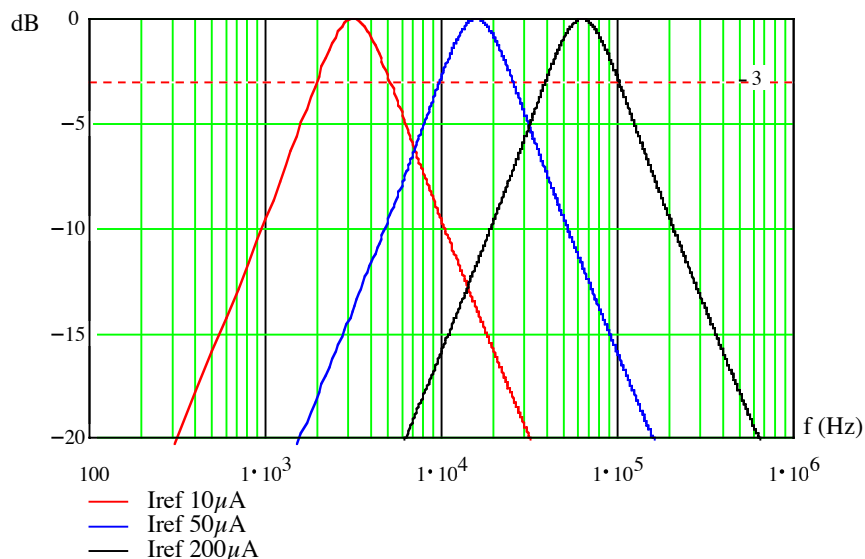
Le graphe asymptotique de Bode du module de cette fonction est identique à celle du premier exercice (Graphe 1 de la figure ci-dessous).

Pour $x \ll 1$ on obtient : $\left| \frac{v_{sl}}{v_e} \right|_{dB} \approx -10 \log \left(1 + \left(\frac{1}{x} \right)^2 \right)$

Le graphe asymptotique de Bode du module de cette fonction correspond au Graphe 2 de la figure ci-dessous. Le graphe passe par x unité avec un coefficient directeur de 20 dB / décade pour $x \rightarrow 0$.



Courbes de réponse en fréquences du filtre passe-bande pour 3 valeurs du courant I_{ref} .



EXERCICE 3 : SORTIE V_{s2} DU FILTRE PASSE-BANDE

2. En exploitant la relation (2) de l'exercice précédent :

$$v_{s2} = \frac{I}{j\omega C(R_1 + R_2)} \frac{g_m R_2}{v_{s1}} \quad \text{avec : } \omega_0 = \frac{g_m R_2}{C(R_1 + R_2)}$$

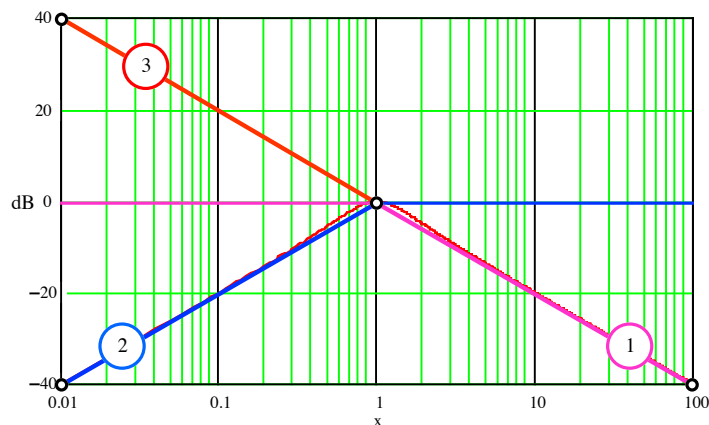
On obtient donc :

$$\frac{v_{s2}}{v_e} = \frac{\frac{I}{jx}}{1 + j(x - \frac{1}{x})}$$

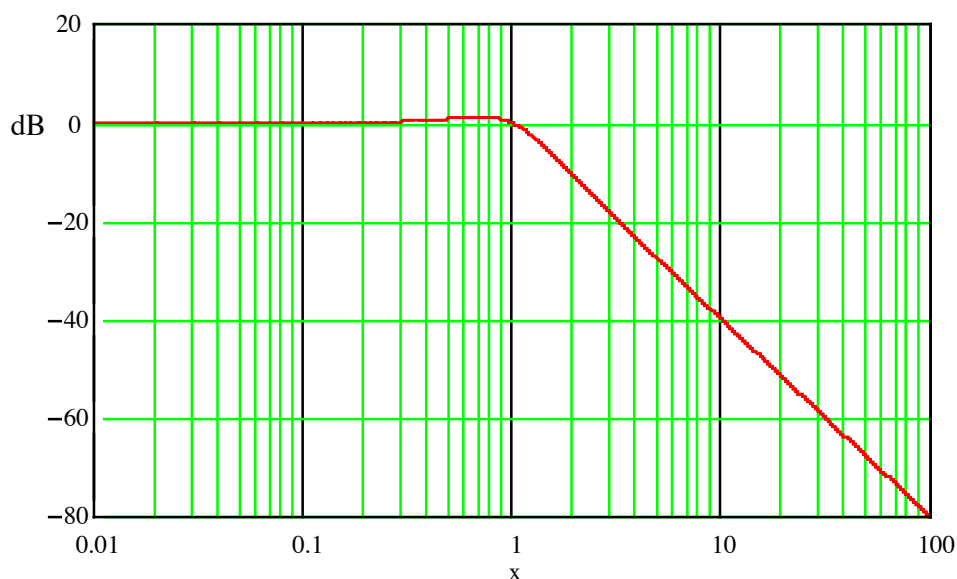
3. Graphe asymptotique de Bode du module de la fonction de transfert du filtre en fonction de la fréquence réduite :

$$\left| \frac{v_{s2}}{v_e} \right|_{dB} = 20 \log\left(\frac{I}{x^2}\right) - 10 \log\left(1 + \left(x - \frac{1}{x}\right)^2\right)$$

Le Graphe de Bode est constitué des courbes (1) et (2) du graphe vu dans l'exercice 2 auquel s'ajoute le tracé (3). Sachant que les graphes (2) et (3) se compensent pour $x < 1$, on obtient finalement un filtre passe-haut au niveau de la sortie v_{s2} .



Courbe de réponse en fréquences du filtre en fonction de x .



LM13600

Dual Operational Transconductance Amplifiers with Linearizing Diodes and Buffers

General Description

The LM13600 series consists of two current controlled transconductance amplifiers each with differential inputs and a push-pull output. The two amplifiers share common supplies but otherwise operate independently. Linearizing diodes are provided at the inputs to reduce distortion and allow higher input levels. The result is a 10 dB signal-to-noise improvement referenced to 0.5 percent THD. Controlled impedance buffers which are especially designed to complement the dynamic range of the amplifiers are provided.

Features

- g_m adjustable over 6 decades
- Excellent g_m linearity

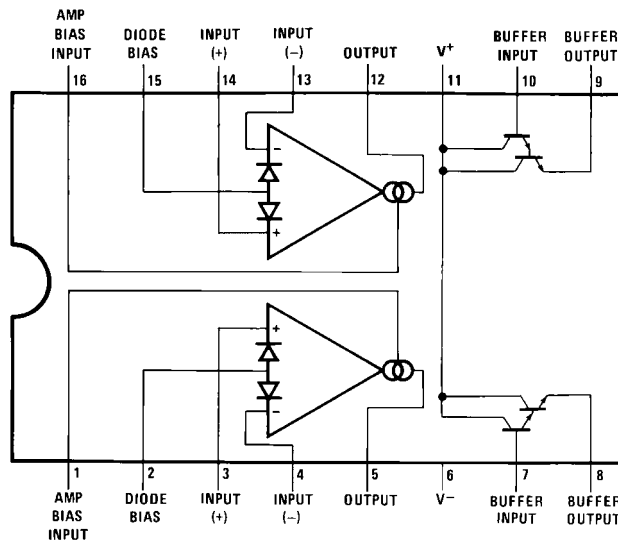
- Excellent matching between amplifiers
- Linearizing diodes
- Controlled impedance buffers
- High output signal-to-noise ratio

Applications

- Current-controlled amplifiers
- Current-controlled impedances
- Current-controlled filters
- Current-controlled oscillators
- Multiplexers
- Timers
- Sample and hold circuits

Connection Diagram

Dual-In-Line and Small Outline Packages



DS007980-2

Top View

Order Number LM13600M, LM13600N or LM13600AN
See NS Package Number M16A or N16A

LM13600 Dual Operational Transconductance Amplifiers with Linearizing Diodes and Buffers

Absolute Maximum Ratings (Note 1)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/Distributors for availability and specifications.

Supply Voltage (Note 2)	
LM13600	36 V _{DC} or ±18V
LM13600A	44 V _{DC} or ±22V
Power Dissipation (Note 3) T _A = 25°C	570 mW
Differential Input Voltage	±5V
Diode Bias Current (I _D)	2 mA
Amplifier Bias Current (I _{ABC})	2 mA
Output Short Circuit Duration	Continuous
Buffer Output Current (Note 4)	20 mA

Operating Temperature Range	0°C to +70°C
DC Input Voltage	+V _S to -V _S
Storage Temperature Range	-65°C to +150°C
Soldering Information	
Dual-In-Line Package	
Soldering (10 seconds)	260°C
Small Outline Package	
Vapor Phase (60 seconds)	215°C
Infrared (15 seconds)	220°C

See AN-450 "Surface Mounting Methods and Their Effect on Product Reliability" for other methods of soldering surface mount devices.

Electrical Characteristics (Note 5)

Parameter	Conditions	LM13600			LM13600A			Units
		Min	Typ	Max	Min	Typ	Max	
Input Offset Voltage (V _{OS})			0.4	4		0.4	1	mV
	Over Specified Temperature Range						2	mV
	I _{ABC} = 5 µA		0.3	4		0.3	1	mV
V _{OS} Including Diodes	Diode Bias Current (I _D) = 500 µA		0.5	5		0.5	2	mV
Input Offset Change	5 µA ≤ I _{ABC} ≤ 500 µA		0.1	3		0.1	1	mV
Input Offset Current			0.1	0.6		0.1	0.6	µA
Input Bias Current			0.4	5		0.4	5	µA
	Over Specified Temperature Range		1	8		1	7	µA
Forward Transconductance (g _m)		6700	9600	13000	7700	9600	12000	µmho
	Over Specified Temperature Range	5400			4000			µmho
g _m Tracking			0.3			0.3		dB
Peak Output Current	R _L = 0, I _{ABC} = 5 µA		5		3	5	7	µA
	R _L = 0, I _{ABC} = 500 µA	350	500	650	350	500	650	µA
	R _L = 0, Over Specified Temp Range	300			300			µA
Peak Output Voltage	Positive		+12	+14.2		+12	+14.2	V
	Negative		-12	-14.4		-12	-14.4	V
Supply Current	I _{ABC} = 500 µA, Both Channels		2.6			2.6		mA
V _{OS} Sensitivity	Positive		20	150		20	150	µV/V
	Negative		20	150		20	150	µV/V
CMRR		80	110		80	110		dB
Common Mode Range		±12	±13.5		±12	±13.5		V
Crosstalk	Referred to Input (Note 6) 20 Hz < f < 20 kHz		100			100		dB
Differential Input Current	I _{ABC} = 0, Input = ±4V		0.02	100		0.02	10	nA
Leakage Current	I _{ABC} = 0 (Refer to Test Circuit)		0.2	100		0.2	5	nA
Input Resistance		10	26		10	26		kΩ
Open Loop Bandwidth			2			2		MHz
Slew Rate	Unity Gain Compensated		50			50		V/µs
Buffer Input Current	(Note 6), Except I _{ABC} = 0 µA		0.2	0.4		0.2	0.4	µA
Peak Buffer Output Voltage	(Note 6)	10			10			V

Electrical Characteristics (Note 5) (Continued)

Note 1: "Absolute Maximum Ratings" indicate limits beyond which damage to the device may occur. Operating Ratings indicate conditions for which the device is functional, but do not guarantee specific performance limits.

Note 2: For selections to a supply voltage above $\pm 22\text{V}$, contact factory.

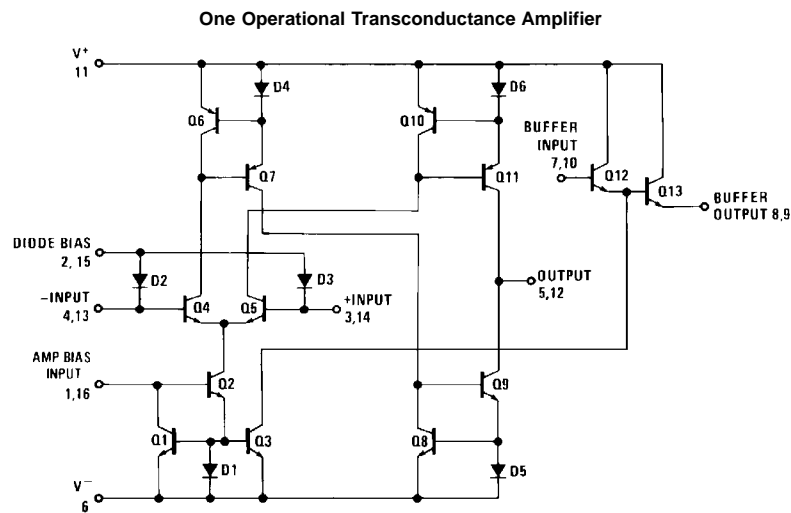
Note 3: For operating at high temperatures, the device must be derated based on a 150°C maximum junction temperature and a thermal resistance of 175°C/W which applies for the device soldered in a printed circuit board, operating in still air.

Note 4: Buffer output current should be limited so as to not exceed package dissipation.

Note 5: These specifications apply for $V_S = \pm 15\text{V}$, $T_A = 25^\circ\text{C}$, amplifier bias current (I_{ABC}) = $500\ \mu\text{A}$, pins 2 and 15 open unless otherwise specified. The inputs to the buffers are grounded and outputs are open.

Note 6: These specifications apply for $V_S = \pm 15\text{V}$, $I_{ABC} = 500\ \mu\text{A}$, $R_{OUT} = 5\ \text{k}\Omega$ connected from the buffer output to $-V_S$ and the input of the buffer is connected to the transconductance amplifier output.

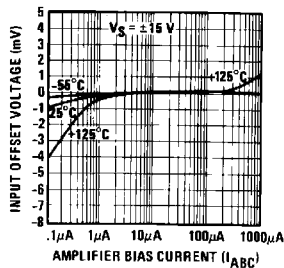
Schematic Diagram



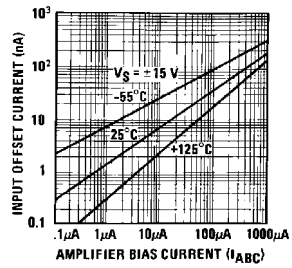
DS007980-1

Typical Performance Characteristics

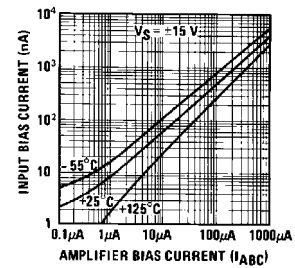
Input Offset Voltage



Input Offset Current

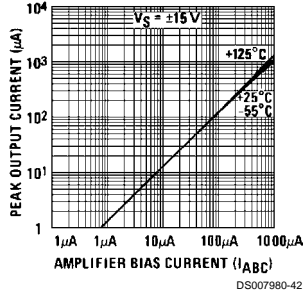


Input Bias Current

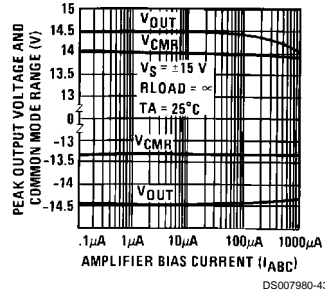


Typical Performance Characteristics (Continued)

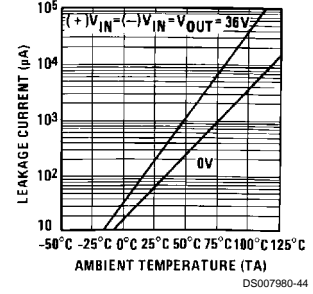
Peak Output Current



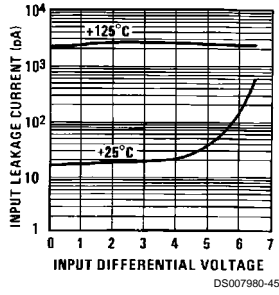
Peak Output Voltage and Common Mode Range



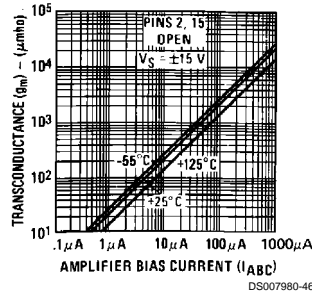
Leakage Current



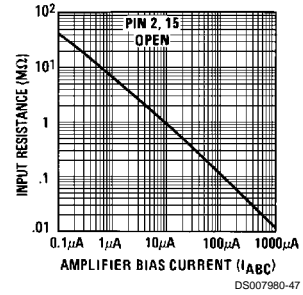
Input Leakage



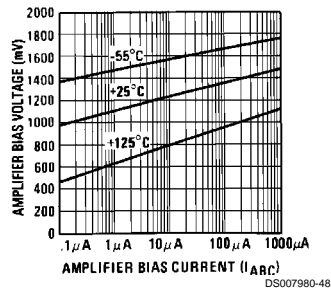
Transconductance



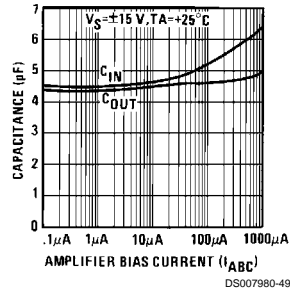
Input Resistance



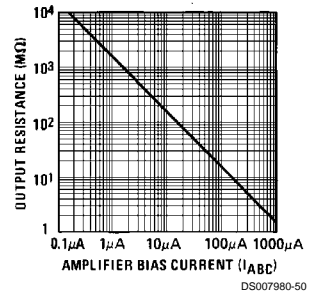
Amplifier Bias Voltage vs Amplifier Bias Current



Input and Output Capacitance

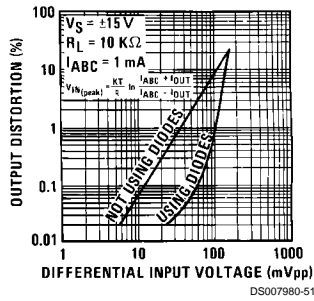


Output Resistance

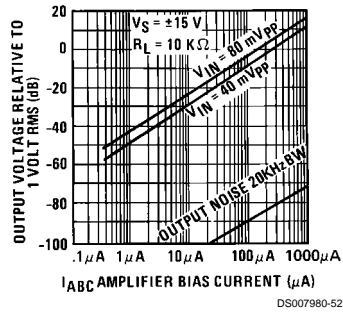


Typical Performance Characteristics (Continued)

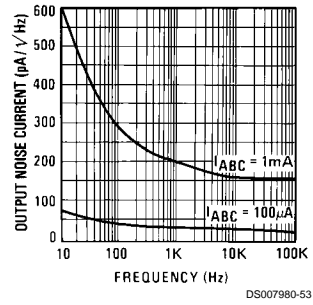
Distortion vs Differential Input Voltage



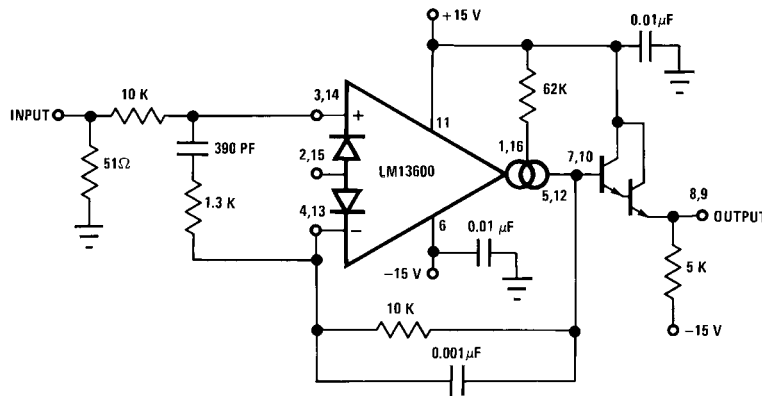
Voltage vs Amplifier Bias Current



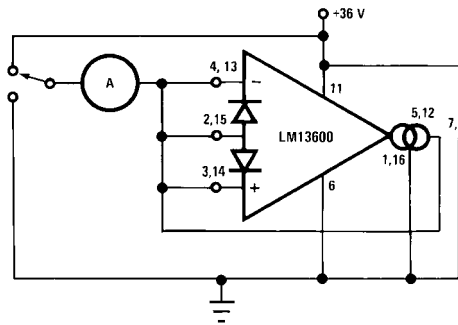
Output Noise vs Frequency



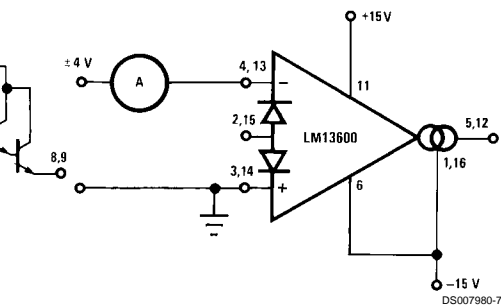
Unity Gain Follower



Leakage Current Test Circuit



Differential Input Current Test Circuit



Circuit Description

The differential transistor pair Q_4 and Q_5 form a transconductance stage in that the ratio of their collector currents is defined by the differential input voltage according to the transfer function:

$$V_{IN} = \frac{kT}{q} \ln \frac{I_5}{I_4} \quad (1)$$

where V_{IN} is the differential input voltage, kT/q is approximately 26 mV at 25°C and I_5 and I_4 are the collector currents of transistors Q_5 and Q_4 respectively. With the exception of Q_3 and $Q_{1,3}$, all transistors and diodes are identical in size. Transistors Q_1 and Q_2 with Diode D_1 form a current mirror which forces the sum of currents I_4 and I_5 to equal I_{ABC} :

$$I_4 + I_5 = I_{ABC} \quad (2)$$

where I_{ABC} is the amplifier bias current applied to the gain pin.

For small differential input voltages the ratio of I_4 and I_5 approaches unity and the Taylor series of the ln function can be approximated as:

$$\begin{aligned} \frac{kT}{q} \ln \frac{I_5}{I_4} &\approx \frac{kT}{q} \frac{I_5 - I_4}{I_4} \\ I_4 &\approx I_5 \approx \frac{I_{ABC}}{2} \end{aligned} \quad (3)$$

$$V_{IN} \left[\frac{I_{ABC} q}{2kT} \right] = I_5 - I_4 \quad (4)$$

Collector currents I_4 and I_5 are not very useful by themselves and it is necessary to subtract one current from the other. The remaining transistors and diodes form three current mirrors that produce an output current equal to I_5 minus I_4 thus:

$$V_{IN} \left[\frac{I_{ABC} q}{2kT} \right] = I_{OUT} \quad (5)$$

The term in brackets is then the transconductance of the amplifier and is proportional to I_{ABC} .

Linearizing Diodes

For differential voltages greater than a few millivolts, Equation (3) becomes less valid and the transconductance becomes increasingly nonlinear. Figure 1 demonstrates how the internal diodes can linearize the transfer function of the amplifier. For convenience assume the diodes are biased with current sources and the input signal is in the form of current I_S . Since the sum of I_4 and I_5 is I_{ABC} and the difference is I_{OUT} , currents I_4 and I_5 can be written as follows:

$$I_4 = \frac{I_{ABC}}{2} - \frac{I_{OUT}}{2}, I_5 = \frac{I_{ABC}}{2} + \frac{I_{OUT}}{2}$$

Since the diodes and the input transistors have identical geometries and are subject to similar voltages and temperatures, the following is true:

$$\begin{aligned} \frac{kT}{q} \ln \frac{\frac{I_D}{2} + I_S}{\frac{I_D}{2} - I_S} &= \frac{kT}{q} \ln \frac{\frac{I_{ABC}}{2} + \frac{I_{OUT}}{2}}{\frac{I_{ABC}}{2} - \frac{I_{OUT}}{2}} \\ \therefore I_{OUT} &= I_S \left(\frac{2I_{ABC}}{I_D} \right) \quad \text{for } |I_S| < \frac{I_D}{2} \end{aligned} \quad (6)$$

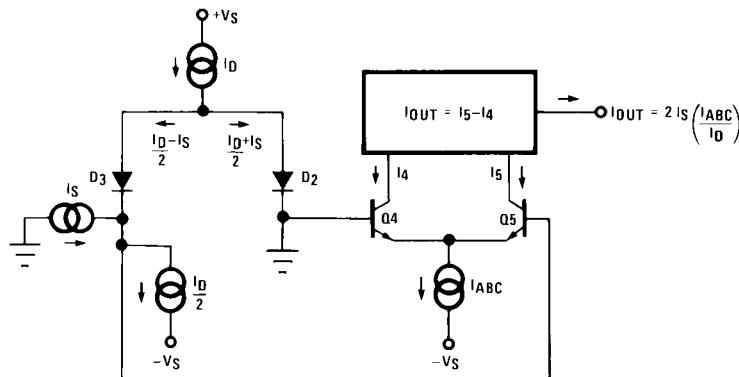


FIGURE 1. Linearizing Diodes

DS007980-8

Linearizing Diodes (Continued)

Notice that in deriving Equation 6 no approximations have been made and there are no temperature-dependent terms. The limitations are that the signal current not exceed $I_D/2$ and that the diodes be biased with currents. In practice, replacing the current sources with resistors will generate insignificant errors.

Controlled Impedance Buffers

The upper limit of transconductance is defined by the maximum value of I_{ABC} (2 mA). The lowest value of I_{ABC} for which the amplifier will function therefore determines the overall dynamic range. At very low values of I_{ABC} , a buffer which has very low input bias current is desirable. An FET follower satisfies the low input current requirement, but is somewhat non-linear for large voltage swing. The controlled impedance buffer is a Darlington which modifies its input bias current to suit the need. For low values of I_{ABC} , the buffer's input current is minimal. At higher levels of I_{ABC} , transistor Q_3 biases up Q_{12} with a current proportional to I_{ABC} for fast slew rate. When I_{ABC} is changed, the DC level of the Darlington output buffer will shift. In audio applications where I_{ABC} is changed suddenly, this shift may produce an audible "pop". For these applications the LM13700 may produce superior results.

Applications-Voltage Controlled Amplifiers

Figure 2 shows how the linearizing diodes can be used in a voltage-controlled amplifier. To understand the input biasing, it is best to consider the 13 k Ω resistor as a current source and use a Thevenin equivalent circuit as shown in Figure 3. This circuit is similar to Figure 1 and operates the same. The potentiometer in Figure 2 is adjusted to minimize the effects of the control signal at the output.

For optimum signal-to-noise performance, I_{ABC} should be as large as possible as shown by the Output Voltage vs. Amplifier Bias Current graph. Larger amplitudes of input signal also improve the S/N ratio. The linearizing diodes help here by allowing larger input signals for the same output distortion as shown by the Distortion vs. Differential Input Voltage graph. S/N may be optimized by adjusting the magnitude of the input signal via R_{IN} (Figure 2) until the output distortion is below some desired level. The output voltage swing can then be set at any level by selecting R_L .

Although the noise contribution of the linearizing diodes is negligible relative to the contribution of the amplifier's internal transistors, I_D should be as large as possible. This minimizes the dynamic junction resistance of the diodes (r_e) and maximizes their linearizing action when balanced against R_{IN} . A value of 1 mA is recommended for I_D unless the specific application demands otherwise.

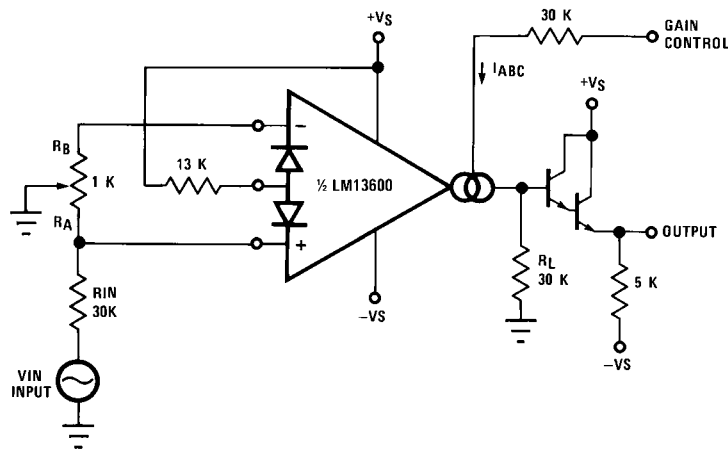
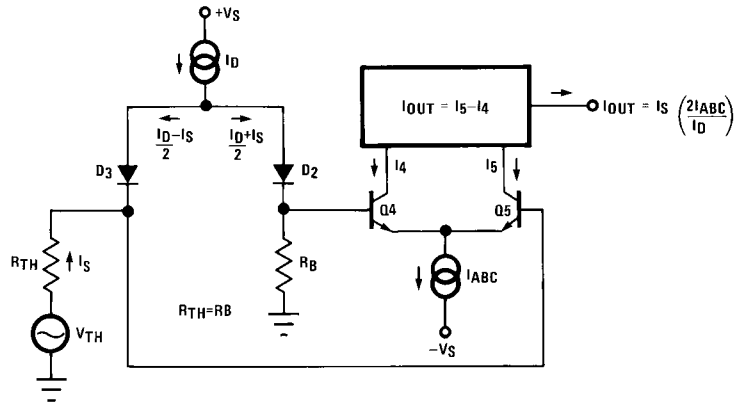


FIGURE 2. Voltage Controlled Amplifier

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Applications-Voltage Controlled Amplifiers (Continued)



DS007980-10

FIGURE 3. Equivalent VCA Input Circuit

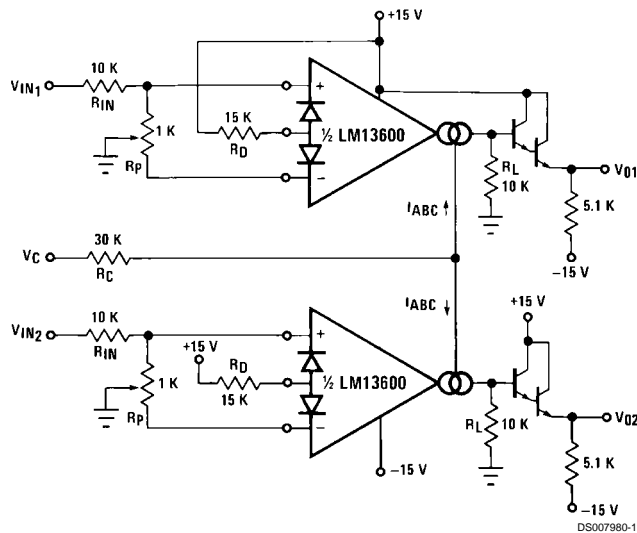
Stereo Volume Control

The circuit of Figure 4 uses the excellent matching of the two LM13600 amplifiers to provide a Stereo Volume Control with a typical channel-to-channel gain tracking of 0.3 dB. R_p is provided to minimize the output offset voltage and may be replaced with two 510Ω resistors in AC-coupled applications. For the component values given, amplifier gain is derived for Figure 2 as being:

$$\frac{V_O}{V_{IN}} = 940 \times I_{ABC}$$

If V_C is derived from a second signal source then the circuit becomes an amplitude modulator or two-quadrant multiplier as shown in Figure 5, where:

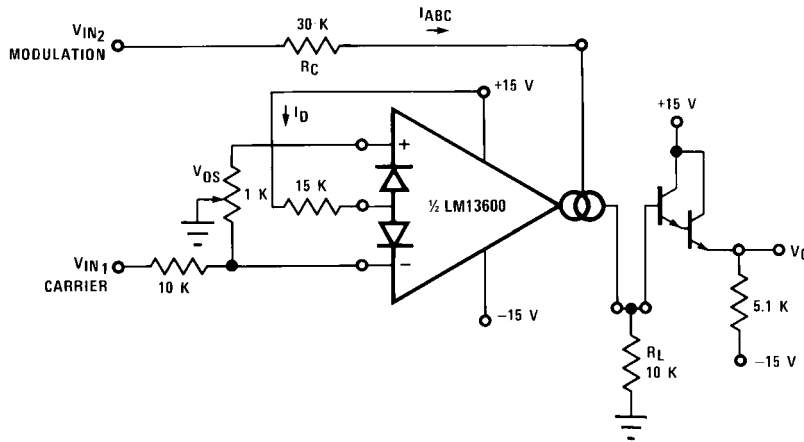
$$I_O = \frac{-2I_S}{I_D} (I_{ABC}) = \frac{-2I_S}{I_D} \frac{V_{IN2}}{R_C} - \frac{2I_S}{I_D} \frac{(V^- + 1.4V)}{R_C}$$



DS007980-11

FIGURE 4. Stereo Volume Control

Stereo Volume Control (Continued)



DS007980-12

FIGURE 5. Amplitude Modulator

The constant term in the above equation can be cancelled by feeding $I_S \times I_D R_C / 2 (V^- + 1.4V)$ into I_O . The circuit of Figure 6 adds R_M to provide this current, resulting in a four-quadrant multiplier where R_C is trimmed such that $V_O = 0V$ for $V_{IN2} = 0V$. R_M also serves as the load resistor for I_O . Noting that the gain of the LM13600 amplifier of Figure 3 may be controlled by varying the linearizing diode current I_D as well as by varying I_{ABC} , Figure 7 shows an AGC Amplifier using this approach. As V_O reaches a high enough amplitude ($3 V_{BE}$) to turn on the Darlington transistors and the linearizing diodes, the increase in I_D reduces the amplifier gain so as to hold V_O at that level.

Voltage Controlled Resistors

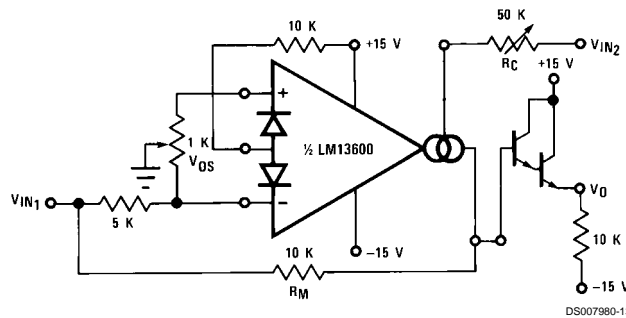
An Operational Transconductance Amplifier (OTA) may be used to implement a Voltage Controlled Resistor as shown in

Figure 8. A signal voltage applied at R_X generates a V_{IN} to the LM13600 which is then multiplied by the g_m of the amplifier to produce an output current, thus:

$$R_X = \frac{R + R_A}{9m R_A}$$

where $g_m \approx 19.2 I_{ABC}$ at $25^\circ C$. Note that the attenuation of V_O by R and R_A is necessary to maintain V_{IN} within the linear range of the LM13600 input.

Figure 9 shows a similar VCR where the linearizing diodes are added, essentially improving the noise performance of the resistor. A floating VCR is shown in Figure 10, where each "end" of the "resistor" may be at any voltage within the output voltage range of the LM13600.



DS007980-13

FIGURE 6. Four-Quadrant Multiplier

Voltage Controlled Resistors (Continued)

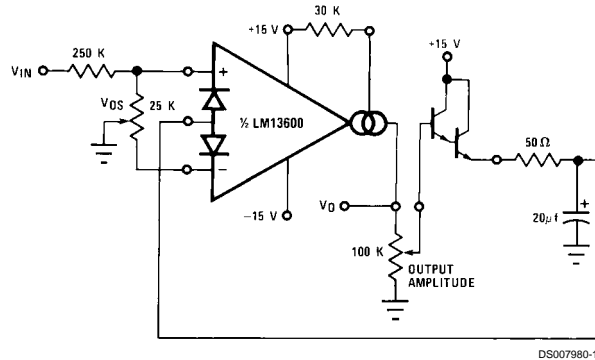


FIGURE 7. AGC Amplifier

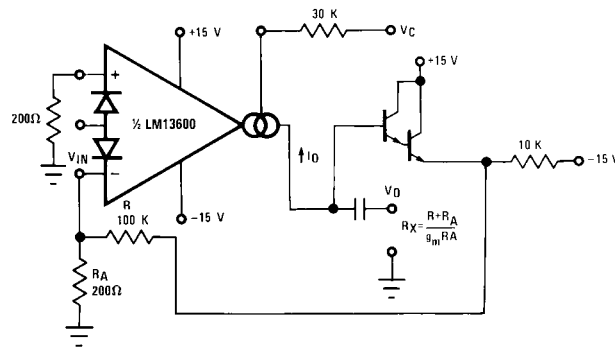


FIGURE 8. Voltage Controlled Resistor, Single-Ended

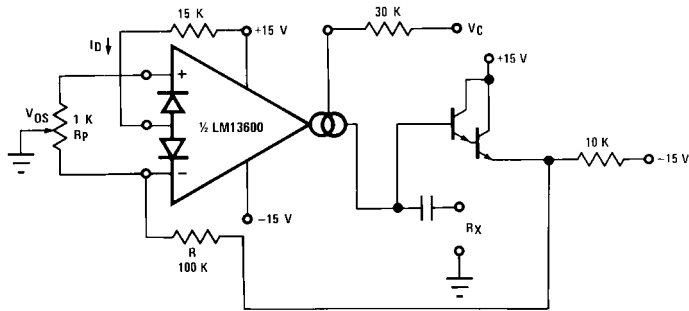
Voltage Controlled Filters

OTA's are extremely useful for implementing voltage controlled filters, with the LM13600 having the advantage that the required buffers are included on the I.C. The VC Lo-Pass Filter of *Figure 11* performs as a unity-gain buffer amplifier at frequencies below cut-off, with the cut-off frequency being the point at which X_C/g_m equals the closed-loop gain of (R/R_A) . At frequencies above cut-off the circuit provides a single RC roll-off (6 dB per octave) of the input signal amplitude with a -3 dB point defined by the given equation, where

g_m is again $19.2 \times I_{ABC}$ at room temperature. *Figure 12* shows a VC High-Pass Filter which operates in much the same manner, providing a single RC roll-off below the defined cut-off frequency.

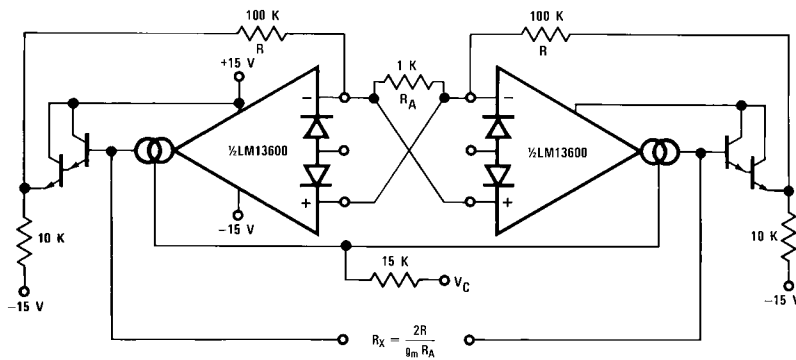
Additional amplifiers may be used to implement higher order filters as demonstrated by the two-pole Butterworth Lo-Pass Filter of *Figure 13* and the state variable filter of *Figure 14*. Due to the excellent g_m tracking of the two amplifiers and the varied bias of the buffer Darlington, these filters perform well over several decades of frequency.

Voltage Controlled Filters (Continued)



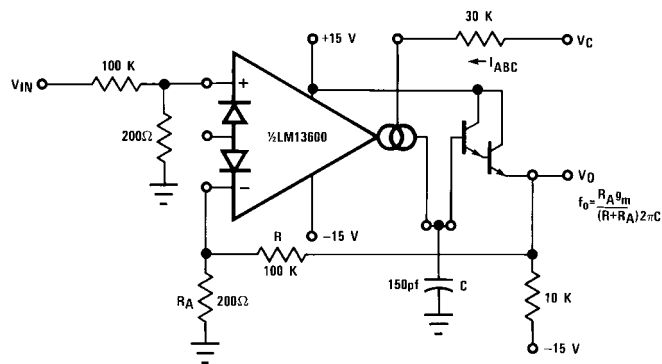
DS007980-16

FIGURE 9. Voltage Controlled Resistor with Linearizing Diodes



DS007980-17

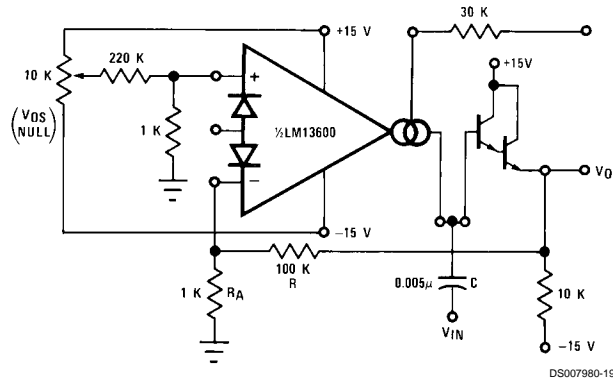
FIGURE 10. Floating Voltage Controlled Resistor



DS007980-18

FIGURE 11. Voltage Controlled Low-Pass Filter

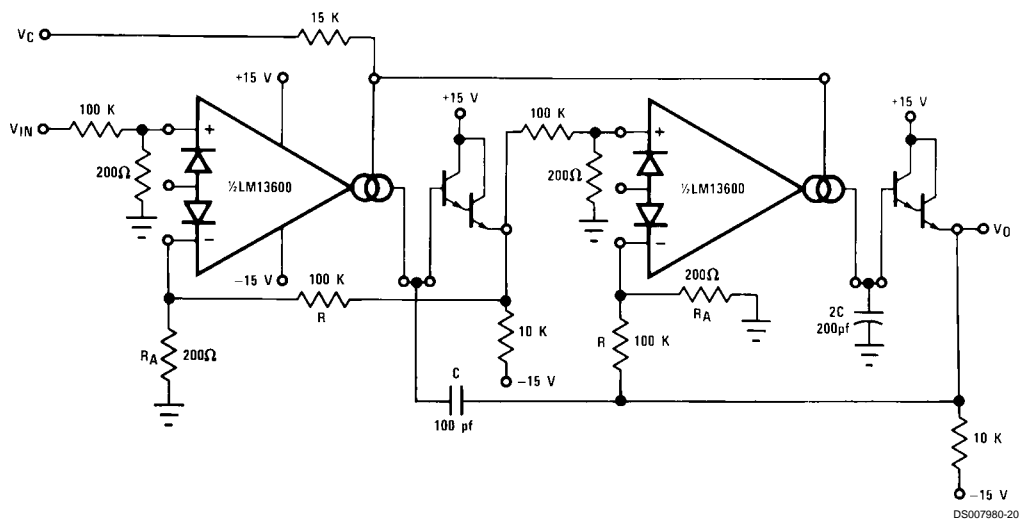
Voltage Controlled Filters (Continued)



DS007980-19

$$f_o = \frac{R_A g_m}{(R + R_A) 2\pi C}$$

FIGURE 12. Voltage Controlled Hi-Pass Filter



DS007980-20

$$f_o = \frac{R_A g_m}{(R + R_A) 2\pi C}$$

FIGURE 13. Voltage Controlled 2-Pole Butterworth Lo-Pass Filter

Voltage Controlled Filters (Continued)

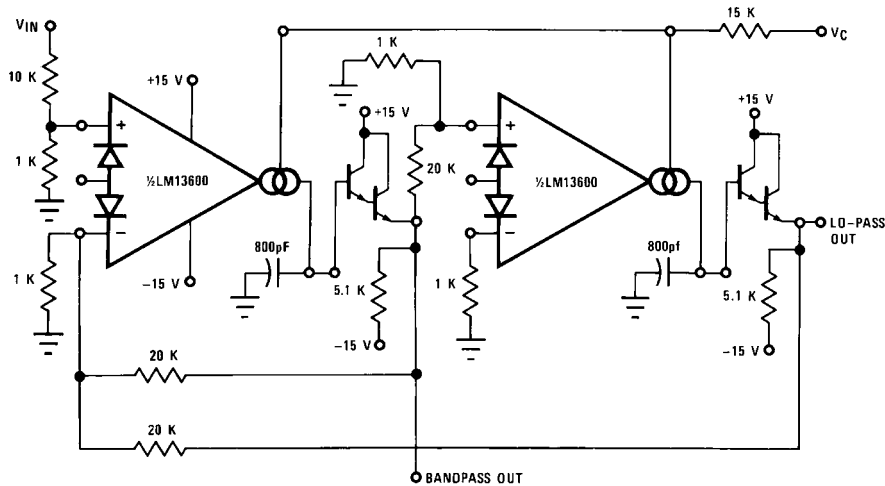


FIGURE 14. Voltage Controlled State Variable Filter

DS007980-21

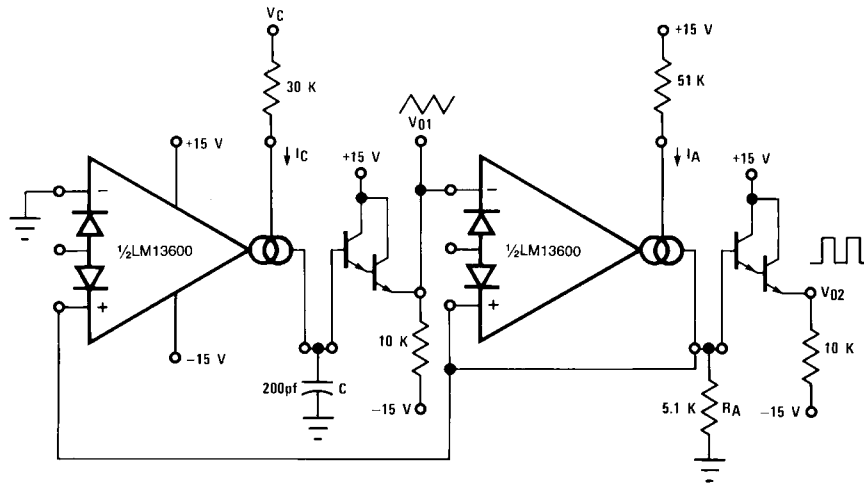
Voltage Controlled Oscillators

The classic Triangular/Square Wave VCO of Figure 15 is one of a variety of Voltage Controlled Oscillators which may be built utilizing the LM13600. With the component values shown, this oscillator provides signals from 200 kHz to below 2 Hz as I_C is varied from 1 mA to 10 nA. The output amplitudes are set by $I_A \times R_A$. Note that the peak differential input voltage must be less than 5V to prevent zenering the inputs.

A few modifications to this circuit produce the ramp/pulse VCO of Figure 16. When V_{O2} is high, I_F is added to I_C to increase amplifier A1's bias current and thus to increase the charging rate of capacitor C. When V_{O2} is low, I_F goes to zero and the capacitor discharge current is set by I_C .

The VC Lo-Pass Filter of Figure 11 may be used to produce a high-quality sinusoidal VCO. The circuit of Figure 16 employs two LM13600 packages, with three of the amplifiers configured as lo-pass filters and the fourth as a limiter/inverter. The circuit oscillates at the frequency at which the loop phase-shift is 360° or 180° for the inverter and 60° per filter stage. This VCO operates from 5 Hz to 50 kHz with less than 1% THD.

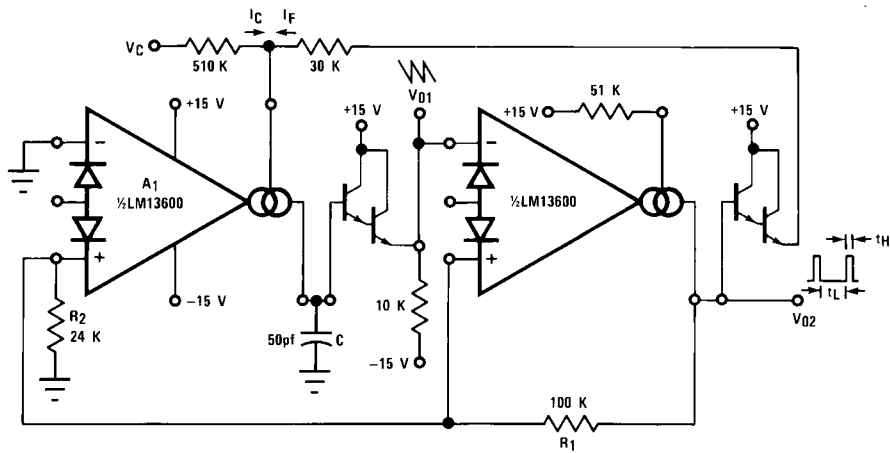
Voltage Controlled Oscillators (Continued)



DS007980-22

$$f_{osc} = \frac{I_C}{4C I_A R_A}$$

FIGURE 15. Triangular/Square-Wave VCO



DS007980-23

$$V_{PK} = \frac{(V^+ - 0.8V)R_2}{R_1 + R_2}$$

$$t_H \approx \frac{2V_{PK}C}{I_F}$$

$$t_L = \frac{2V_{PK}C}{I_C}$$

$$f_o \approx \frac{I_C}{2V_{PK}C} \text{ for } I_C \ll I_F$$

FIGURE 16. Ramp/Pulse VCO

Voltage Controlled Oscillators (Continued)

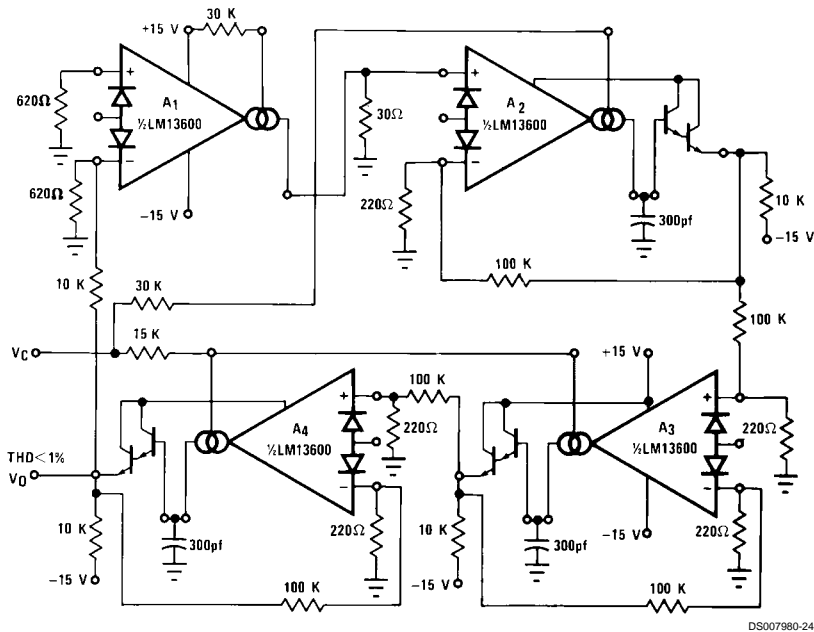


FIGURE 17. Sinusoidal VCO

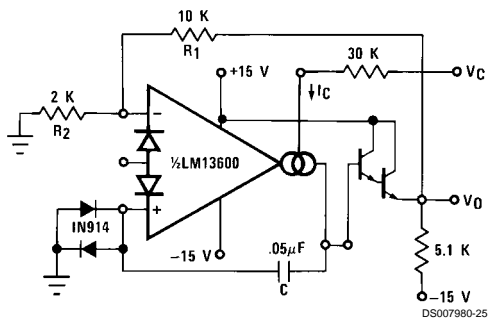


FIGURE 18. Single Amplifier VCO

Figure 18 shows how to build a VCO using one amplifier when the other amplifier is needed for another function.

Additional Applications

Figure 19 presents an interesting one-shot which draws no power supply current until it is triggered. A positive-going trigger pulse of at least 2V amplitude turns on the amplifier through R_B and pulls the non-inverting input high. The amplifier regenerates and latches its output high until capacitor C charges to the voltage level on the non-inverting input. The output then switches low, turning off the amplifier and discharging the capacitor. The capacitor discharge rate is increased by shorting the diode bias pin to the inverting input so that an additional discharge current flows through D₁

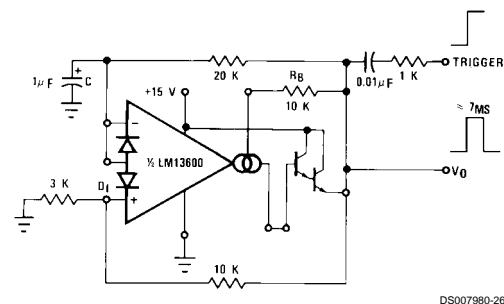


FIGURE 19. Zero Stand-By Power Timer

Additional Applications (Continued)

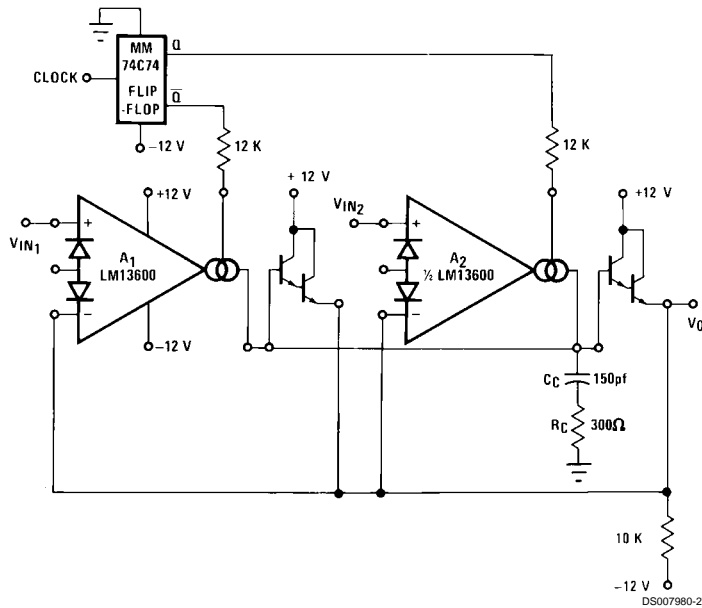


FIGURE 20. Multiplexer

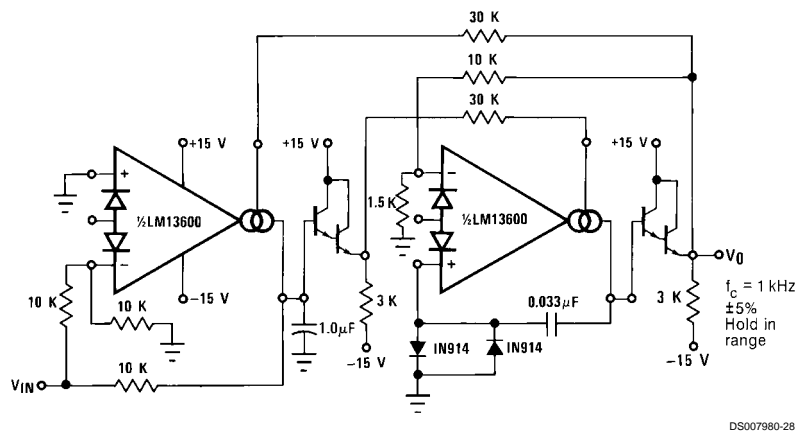


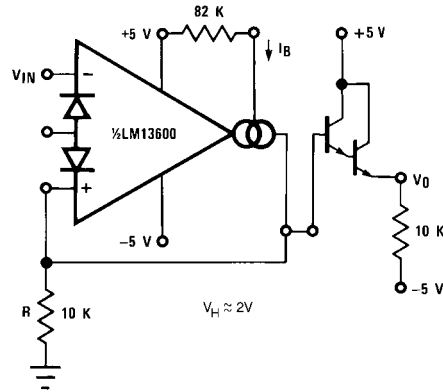
FIGURE 21. Phase Lock Loop

The Schmitt Trigger of *Figure 22* uses the amplifier output current into R to set the hysteresis of the comparator; thus $V_H = 2 \times R \times I_B$. Varying I_B will produce a Schmitt Trigger with variable hysteresis.

Figure 23 shows a Tachometer or Frequency-to-Voltage converter. Whenever A1 is toggled by a positive-going input, an amount of charge equal to $(V_H - V_L) C_t$ is sourced into C_t and R_t . This once-per-cycle charge is then balanced by the current of V_O/R_t . The maximum f_{IN} is limited by the amount of time required to charge C_t from V_L to V_H with a current of I_B , where V_L and V_H represent the maximum low and maximum high output voltage swing of the LM13600. D1 is added to provide a discharge path for C_t when A1 switches low.

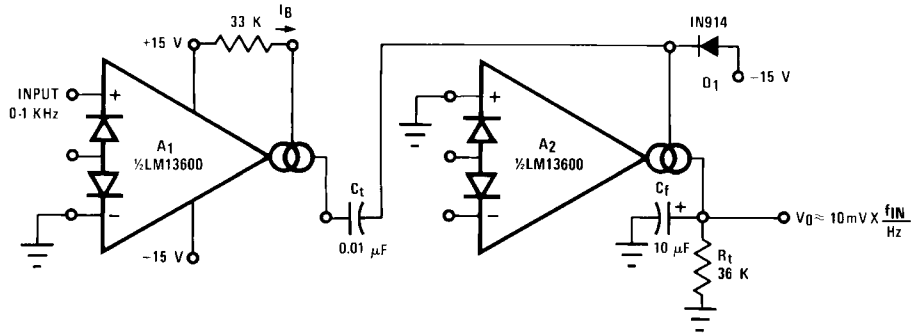
The Peak Detector of *Figure 24* uses A2 to turn on A1 whenever V_{IN} becomes more positive than V_O . A1 then charges storage capacitor C to hold V_O equal to V_{INPK} . One precaution to observe when using this circuit: the Darlington transistor used must be on the same side of the package as A2 since the A1 Darlington will be turned on and off with A1. Pulling the output of A2 low through D1 serves to turn off A1 so that V_O remains constant.

Additional Applications (Continued)



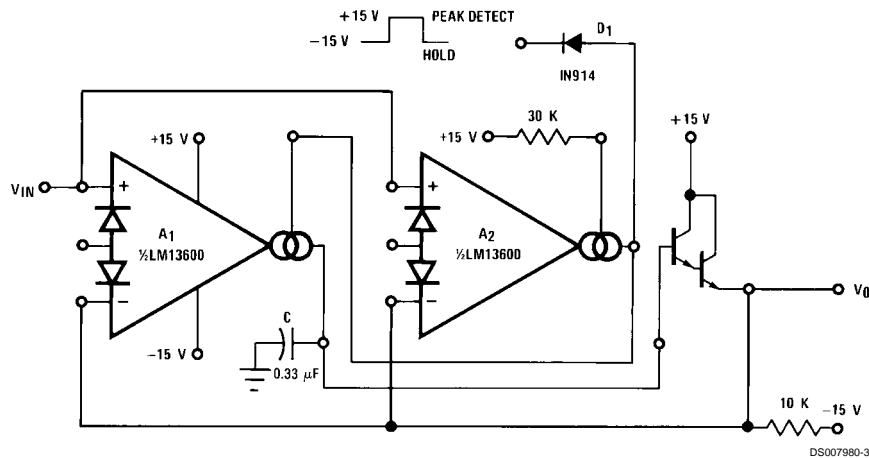
DS007980-29

FIGURE 22. Schmitt Trigger



DS007980-30

FIGURE 23. Tachometer



DS007980-31

FIGURE 24. Peak Detector and Hold Circuit

The Sample-and-Hold circuit of Figure 25 also requires that the Darlington buffer used be from the other (A2) half of the package and that the corresponding amplifier be biased on

continuously. The Ramp-and-Hold of Figure 26 sources I_B

Additional Applications (Continued)

into capacitor C whenever the input to A1 is brought high, giving a ramp-rate of about 1 V/ms for the component values shown.

The true-RMS converter of *Figure 27* is essentially an automatic gain control amplifier which adjusts its gain such that the AC power at the output of amplifier A1 is constant. The output power of amplifier A1 is monitored by squaring amplifier A2 and the average compared to a reference voltage with amplifier A3. The output of A3 provides bias current to the diodes of A1 to attenuate the input signal. Because the output power of A1 is held constant, the RMS value is constant and the attenuation is directly proportional to the RMS value of the input voltage. The attenuation is also proportional to the diode bias current. Amplifier A4 adjusts the ratio of currents through the diodes to be equal and therefore the voltage at the output of A4 is proportional to the RMS value of the input voltage. The calibration potentiometer is set such that V_O reads directly in RMS volts.

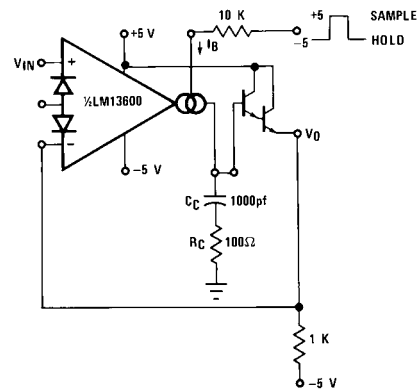


FIGURE 25. Sample-Hold Circuit

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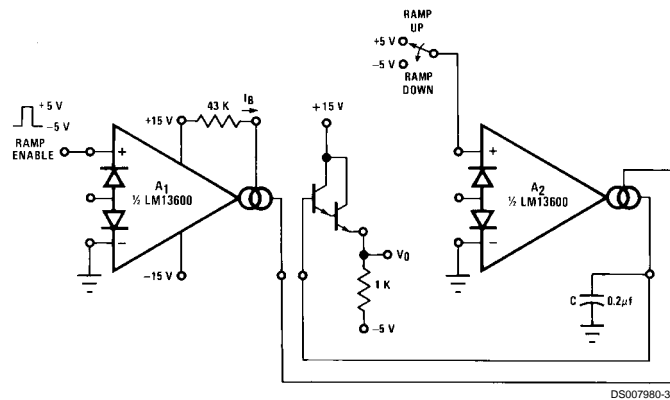


FIGURE 26. Ramp and Hold

DS007980-33

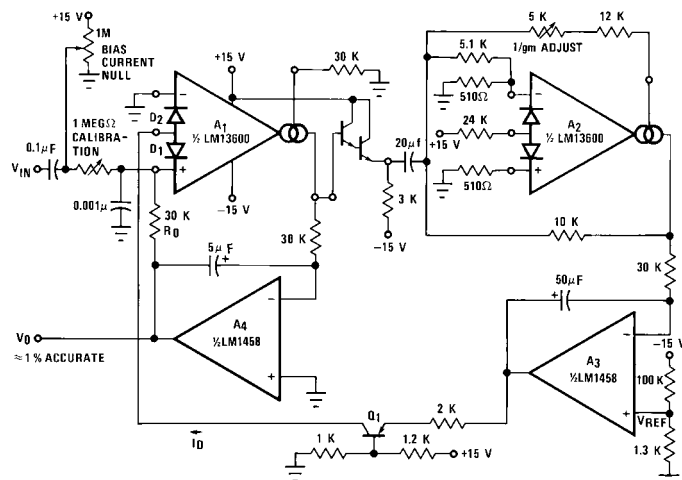


FIGURE 27. True RMS Converter

DS007980-34

Additional Applications (Continued)

The circuit of *Figure 28* is a voltage reference of variable temperature coefficient. The 100 kΩ potentiometer adjusts the output voltage which has a positive TC above 1.2V, zero TC at about 1.2V and negative TC below 1.2V. This is accomplished by balancing the TC of the A2 transfer function against the complementary TC of D1.

The log amplifier of *Figure 29* responds to the ratio of currents through buffer transistors Q3 and Q4. Zero temperature dependence for V_{OUT} is ensured because the TC of the A2 transfer function is equal and opposite to the TC of the logging transistors Q3 and Q4.

The wide dynamic range of the LM13600 allows easy control of the output pulse width in the Pulse Width Modulator of *Figure 30*.

For generating I_{ABC} over a range of 4 to 6 decades of current, the system of *Figure 31* provides a logarithmic current out for a linear voltage in.

Since the closed-loop configuration ensures that the input to A2 is held equal to 0V, the output current of A1 is equal to $I_3 = -V_C/R_C$.

The differential voltage between Q1 and Q2 is attenuated by the R1, R2 network so that A1 may be assumed to be operating within its linear range. From *Equation (5)*, the input voltage to A1 is:

$$V_{IN1} = \frac{-2kT I_3}{q I_2} = \frac{2kT V_C}{q I_2 R_C}$$

The voltage on the base of Q1 is then

$$V_{B1} = \frac{(R_1 + R_2) V_{IN1}}{R_1}$$

The ratio of the Q1 and Q2 collector currents is defined by:

$$V_{B1} = \frac{kT}{q} \ln \frac{I_{C2}}{I_{C1}} \approx \frac{kT}{q} \ln \frac{I_{ABC}}{I_1}$$

Combining and solving for I_{ABC} yields:

$$I_{ABC} = I_1 \exp \left[\frac{2(R_1 + R_2) V_C}{R_1 I_2 R_C} \right]$$

This logarithmic current can be used to bias the circuit of *Figure 4* provide a temperature independent stereo attenuation characteristic.

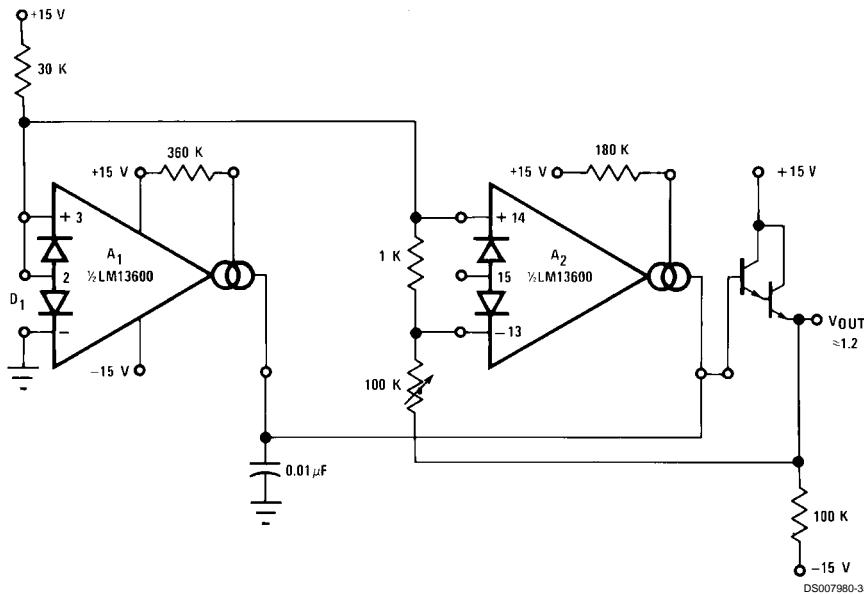
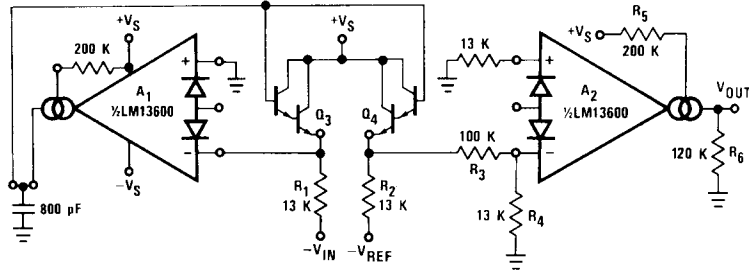


FIGURE 28. Delta VBE Reference

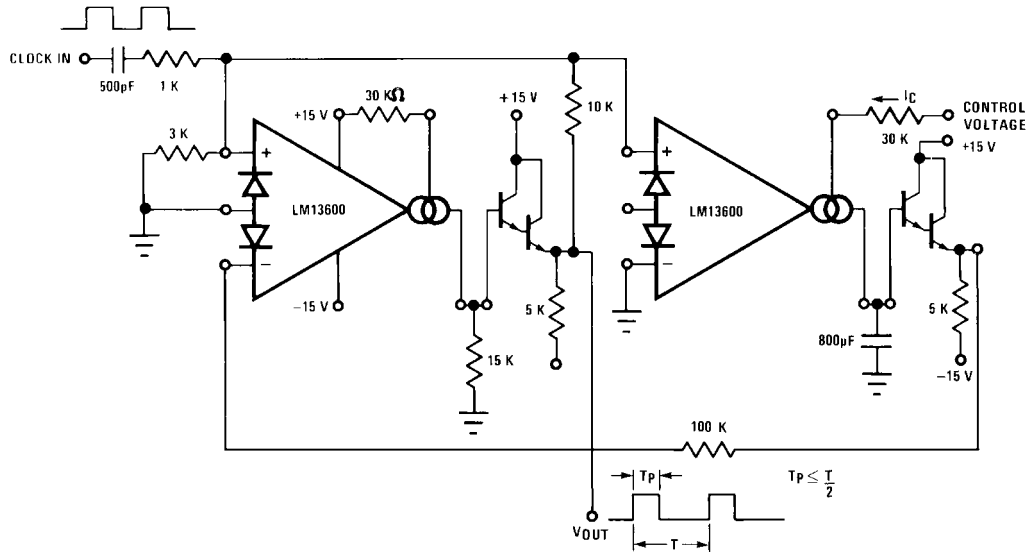
Additional Applications (Continued)



DS007980-36

$$V_{OUT} = \frac{(2V_S - 1.2V)(R_4)(R_6)}{(R_3 + R_4)(R_5)} \ln \frac{V_{IN} R_2}{V_{REF} R_1}$$

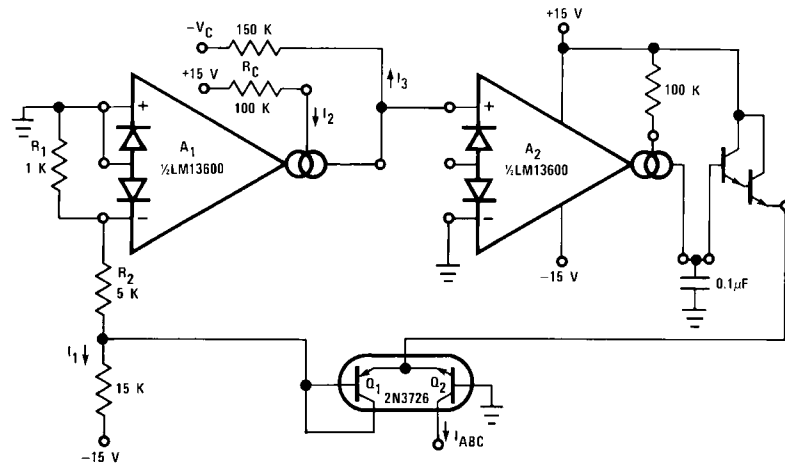
FIGURE 29. Log Amplifier



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FIGURE 30. Pulse Width Modulator

Additional Applications (Continued)



DS007980-38

$$I_{ABC} = I_1 \exp \left[\frac{-CI_3}{I_2} \right]$$

FIGURE 31. Logarithmic Current Source