SECTION V

HIGH SPEED NON-LINEAR SIGNAL PROCESSORS:

MULTIPLIERS, MODULATORS & LOG AMPS

HIGH SPEED NON-LINEAR SIGNAL PROCESSORS:

MULTIPLIERS, MODULATORS & LOG AMPS

INTRODUCTION

MULTIPLIERS

MODULATORS

LOGARITHMIC AMPLIFIERS

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INTRODUCTION

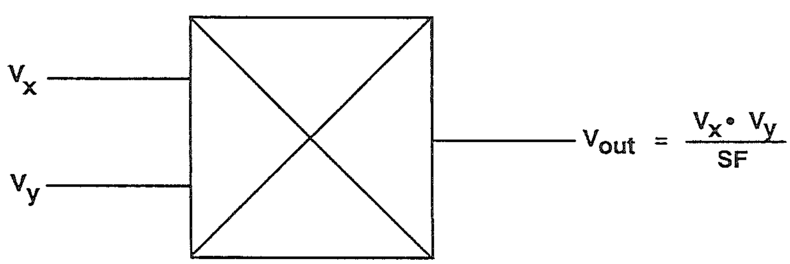
Historically analog computers have been slow devices. Although high frequency multipliers, modulators, logarithmic amplifiers and other function generators have been available for many years they have generally had relatively poor accuracy and stability and have not usually been considered as "analog computers". This section of our seminar considers a few classes of non-linear device which perform accurate analog operations of various sorts at high speeds. These devices are multipliers and modulators, and logarithmic amplifiers.

MULTIPLIERS

A multiplier is a device having two input ports and an output port. The signal at the output is the product of the two input signals. If both input and output signals are voltages, the transfer characteristic is

$$V_{out} = \frac{V_x \bullet V_y}{SF}$$, where

SF is a scaling factor and, in order that the equation be dimensionally correct, has the dimension of voltage. (See Figure 5.1)



BASIC MULTIPLIER

Figure 5.1

From a mathematical point of view multiplication is a "four quadrant" operation - that is to say that both V_x and V_y may be either positive or negative, as may be V_{out}. Some of the circuits used to produce electronic multipliers, however, are limited to signals of one polarity.

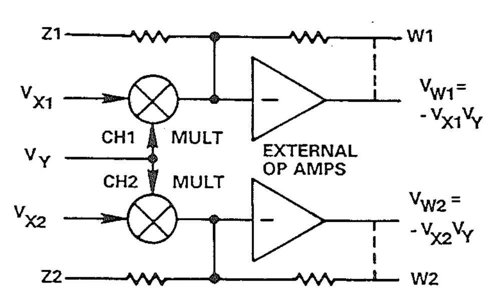
If both signals must be unipolar we have a "single quadrant" multiplier and the output will also be unipolar. If one of the signals is unipolar but the other may have either polarity, then the multiplier is a "two quadrant" multiplier and the output may have either polarity (and is "bipolar").

TYPES OF MULTIPLIERS

TYPE	V_X	V_Y	V_{OUT}
Single quadrant	Unipolar	Unipolar	Unipolar
Two quadrant	Bipolar	Unipolar	Bipolar
Four quadrant	Bipolar	Bipolar	Bipolar

Figure 5. 2

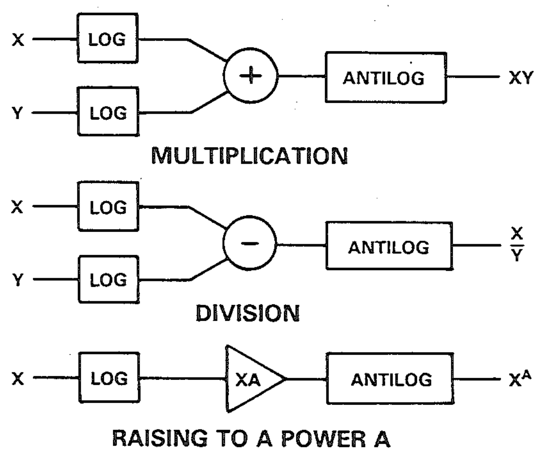
The circuitry used to produce one- and two-quadrant multipliers may be simpler than that required for four quadrant multipliers, and since there are many applications where full four quadrant multiplication is not required, it is by no means uncommon to find accurate devices which work in only one or two quadrants. An example is the AD539, which is a wideband dual two-quadrant multiplier which has a single unipolar V_y input with the relatively limited bandwidth of 5 MHz, and two bipolar V_x inputs, one per multiplier, with bandwidths of 60 MHz. A block diagram of the AD539 is shown in Figure 5.3.



AD539 FUNCTIONAL BLOCK DIAGRAM

Figure 5.3

The simplest electronic multiplier circuits use logarithmic amplifiers. The electronic principles involved will be described later, but the computation relies on the simple fact that the antilog of the sum of the logs of two numbers is the product of those numbers. (See Figure 5.4.)



COMPUTATION WITH LOG AND ANTILOG CIRCUITS

Figure 5.4

The disadvantages of this type of multiplication are a very limited bandwidth and single quadrant operation. A far better type of multiplier uses the "Gilbert Cell". This structure was invented by Barrie Gilbert in the late 1960s.[1, 2]

There is a linear relationship between the collector current of a silicon junction transistor and its transconductance (gain) which is given by

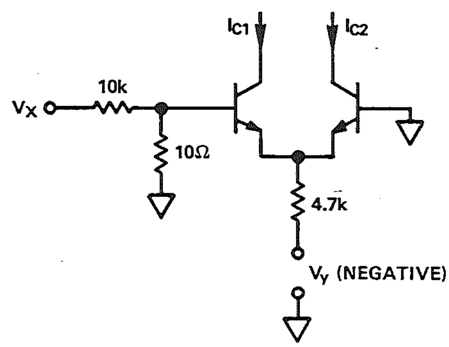
$$\frac{dI_c}{dV_{be}} = \frac{qI_c}{kT}$$

Where

I_c = the collector current
V_{be} = the base-emitter voltage
q = the electron charge (1.60219E-19)
k = Boltzmann's constant (1.38062E-23)
T = the absolute temperature

This relationship may be exploited to construct a multiplier with a long-tailed pair of silicon transistors, as shown in Figure 5.5.

This is rather a poor multiplier because (1) the Y input is offset by the V_{be} - which changes non-linearly with V_y; (2) the X input is non-linear as a result of the exponential relationship between I_c and V_{be}; and (3) the scale factor varies with temperature.



$$I_{C1} - I_{C2} = \Delta I_C = \frac{q}{kT}\left(\frac{V_y + V_{be}}{4.7 \times 10^3}\right)\left(\frac{10}{10{,}010}\right)V_x$$

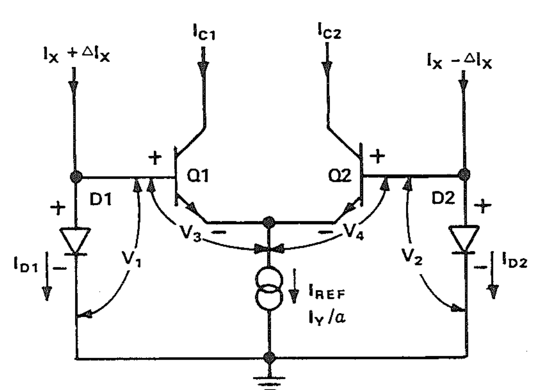
$$= 8.3 \times 10^{-6}\ (V_y + 0.6)\ V_x \quad @\ 25°C$$

BASIC TRANSCONDUCTANCE MULTIPLIER CIRCUIT

Figure 5.5

Gilbert realized that this circuit could be linearized and made temperature stable by working with currents, rather than voltages, and by exploiting the logarithmic I_C/V_{be} properties of transistors. (See Figure 5.6.)

$$\Delta I_C = I_{C1} - I_{C2}$$



$$\Delta I_C = \frac{\Delta I_X I_Y}{I_X}$$

THE GILBERT CELL - A LINEAR TWO-QUADRANT MULTIPLIER

Figure 5.6

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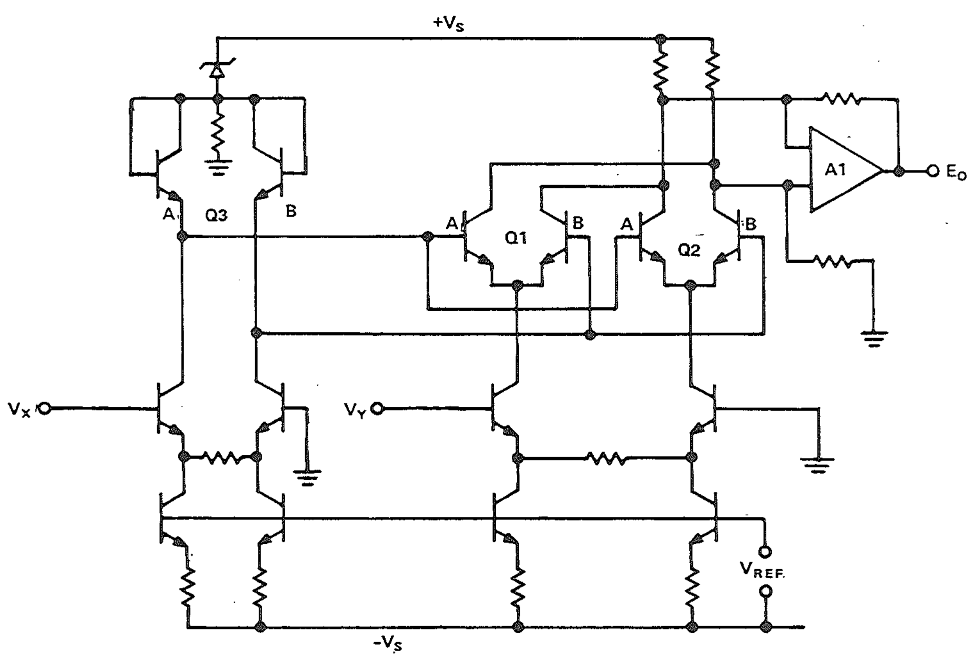
The X input to the Gilbert Cell takes the form of a differential current, and the Y input is a unipolar current. The differential X currents flow in two diode connected transistors and the logarithmic voltages compensate for the exponential V_{be}/I_c relationship. Furthermore the q/kT scale factors cancel. This gives the Gilbert Cell the linear transfer function

$$\Delta I_c = \frac{\Delta I_x \bullet I_y}{I_x}$$

As it stands the Gilbert Cell has three inconvenient features:

(1) Its X input is a differential current;
(2) Its output is a differential current; and
(3) Its Y input is a unipolar current - so the cell is only a two quadrant multiplier.

By cross-coupling two such cells and using two voltage-to-current converters (as shown in Figure 5.7) , we can convert the basic architecture to a four quadrant device with voltage inputs. At low and medium frequencies a subtractor amplifier may be used to convert the differential current at the output to a voltage - but this may not be the best way of handling signals at high frequencies.



4-QUADRANT TRANSLINEAR MULTIPLIER

Figure 5.7

In Figure 5. 7, Q1A & Q1B, and Q2A & Q2B form the two core long-tailed pairs of the two Gilbert Cells, while Q3A and Q3B are the linearizing transistors for both cells. In Figure 5.7 there is an operational amplifier acting as a differential current to single-ended voltage converter, but for higher speed applications the cross-coupled collectors of Q1 and Q2 might form a differential open collector current output.

This multiplier relies on the matching of a number of transistors and currents. This is easily accomplished on a monolithic chip but is much harder with discrete components - such multipliers are, therefore, almost invariably made as ICs. Even the best integrated circuit processes have some residual errors, however, and these show up as four DC errors in such multipliers.

TRIMMABLE ERRORS IN MULTIPLIERS

X-Input Voltage Offset:	Y Feedthrough
Y-Input Voltage Offset:	X Feedthrough
Z-Input (output amplifier) Voltage Offset:	dc output offset voltage
Resistor Mismatch:	Gain Error

Figure 5.8

In early Gilbert Cell multipliers these errors had to be trimmed by means of resistors and potentiometers external to the chip, which was somewhat inconvenient. With modern analog processes, which permit the laser trimming of SiCr thin film resistors on the chip itself, it is possible to trim these errors during manufacture so that the final device has very high accuracy. Internal trimming has the additional advantage that it does not reduce the high frequency performance, as may be the case with external trimpots.

KEY FEATURES OF THE TRANSLINEAR MULTIPLIER

- High Accuracy - better than 0.1% is posssible
- Wide Bandwidth (over 60 MHz voltage output over 500 MHz current output)
- Simplicity, low cost, and ease of use

Figure 5.9

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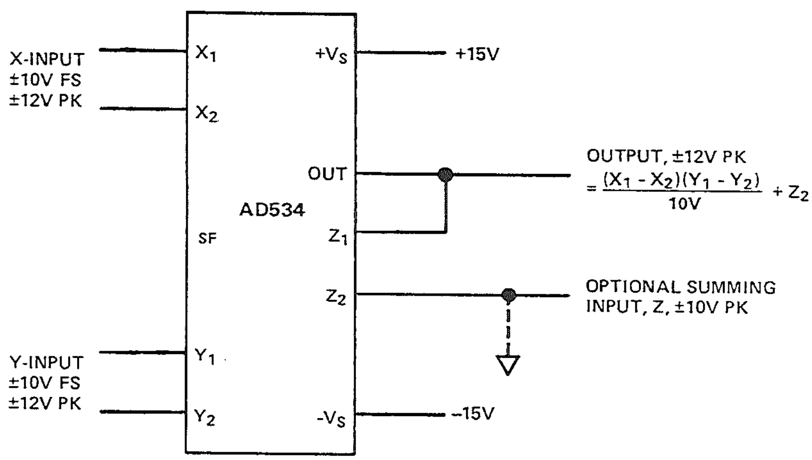
Because the internal structure of the translinear multiplier is necessarily differential, the inputs are usually differential as well (after all, if a single-ended input is required it is not hard to ground one of the inputs). This is not only convenient in allowing common-mode signals to be rejected, it also permits more complex computations to be performed.

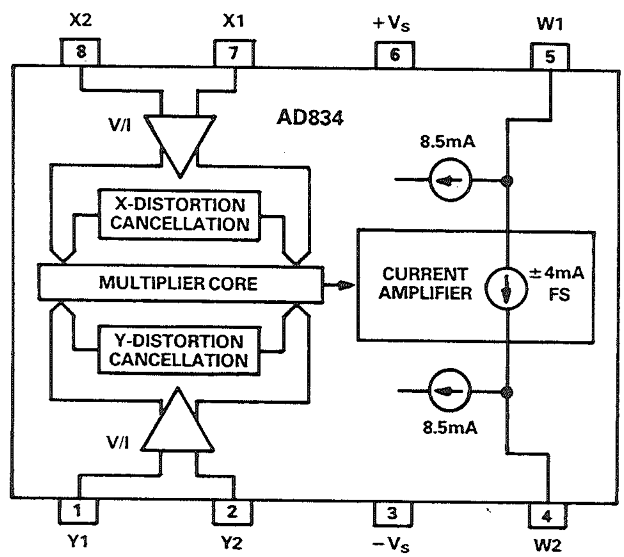
The AD534 (see Figure 5.10) is the classic example of a four-quadrant multiplier based on the Gilbert Cell. It has an accuracy of 0.1% in the multiplier mode, fully differential inputs, and a voltage output. However, as a result of its voltage output architecture, its bandwidth is only about 1MHz.



THE AD534 MULTIPLIER

Figure 5.10

For wideband applications the basic multiplier with open collector current outputs is used. The AD834 is an 8-pin device with differential X inputs, differential Y inputs, differential open collector current outputs, and a bandwidth of over 500 MHz. A block diagram is shown in Figure 5.11.



THE AD834 FUNCTIONAL BLOCK DIAGRAM

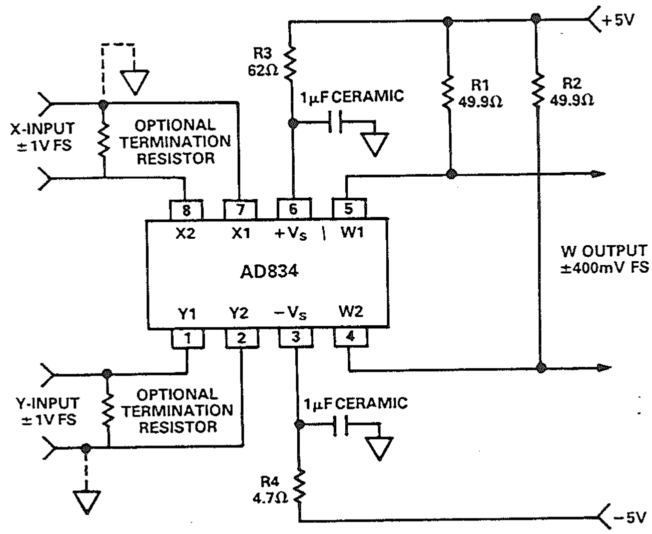
Figure 5.11

The AD834 is a true linear multiplier with a transfer function of

$$I_{out} = \frac{V_x \bullet V_y}{1V \bullet 250\Omega}$$

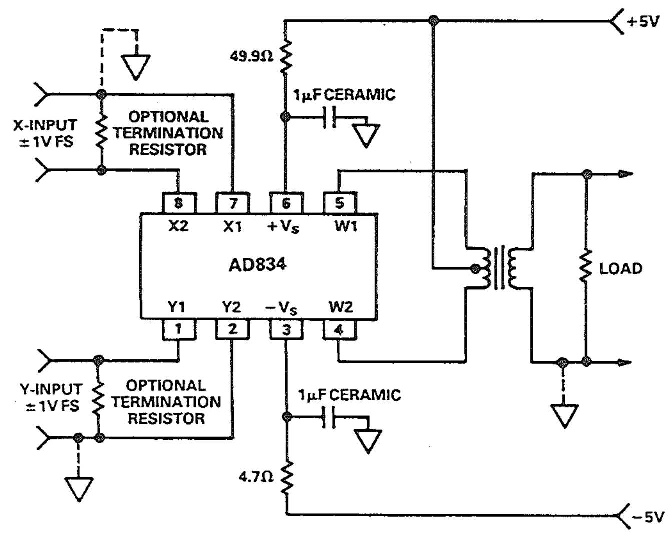
Its X and Y offsets are trimmed to 500 μV (3 mV max), and it may be used in a wide variety of applications including multipliers (broadband and narrowband), squarers, frequency doublers, and high frequency power measurement circuits. A consideration when using the AD834 is that, because of its very wide bandwidth, its input bias currents, approx 50 μA per input, must be considered in the design of input circuitry lest, flowing in source resistances, they give rise to unplanned offset voltages.

A basic wideband multiplier using the AD834 is shown in Figure 5.12. The differential current output flows in equal load resistors R1 and R2 to give a differential voltage output. This is the simplest application circuit for the device. Where only the high frequency outputs are required, transformer coupling may be used, with either simple transformers or, for better wideband performance, transmission line or "Ruthroff"[3] transformers (see Figure 5.13).



BASIC CONNECTIONS FOR WIDEBAND AD834 OPERATION

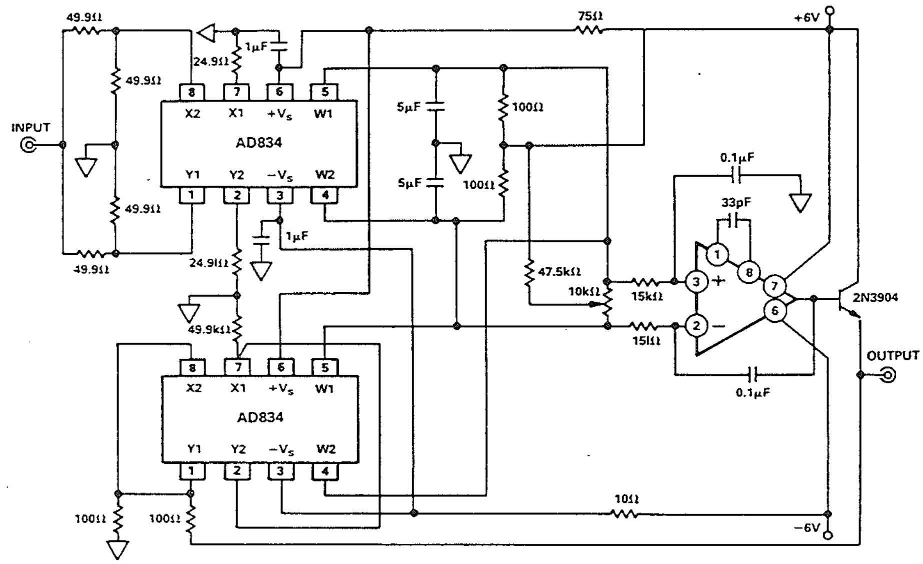
Figure 5.12



AD834 OPERATION WITH WIDEBAND TRANSFORMER COUPLED OUTPUT

Figure 5.13

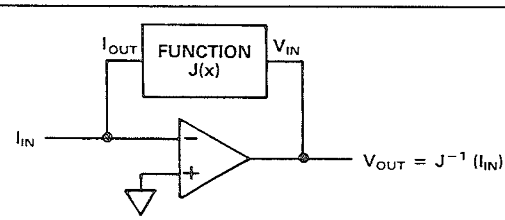
A very wideband (500 MHz) root mean square (rms) circuit is shown in Figure 5.14. This uses two AD834s, one as a wideband squarer, and one as a square rooter.



WIDEBAND R.M.S. MEASUREMENT WITH THE AD834

Figure 5.14

The operation of the circuit is described in detail in the data sheet of the AD834, but in essence the signal is applied to both inputs of the top AD834 whose output is then the square of the instantaneous value of the input. This output is averaged in the two 5 μF capacitors, and the square root of this mean square is computed by the op-amp and the second AD834, using the common principle of analog computing that a function generator in a negative feedback loop computes the inverse function (provided, of course, that the function is monotonic over the range of operation). (See Figure 5.15.)



NOTE: FUNCTION MUST BE MONOTONIC
OVER THE RELEVANT RANGE

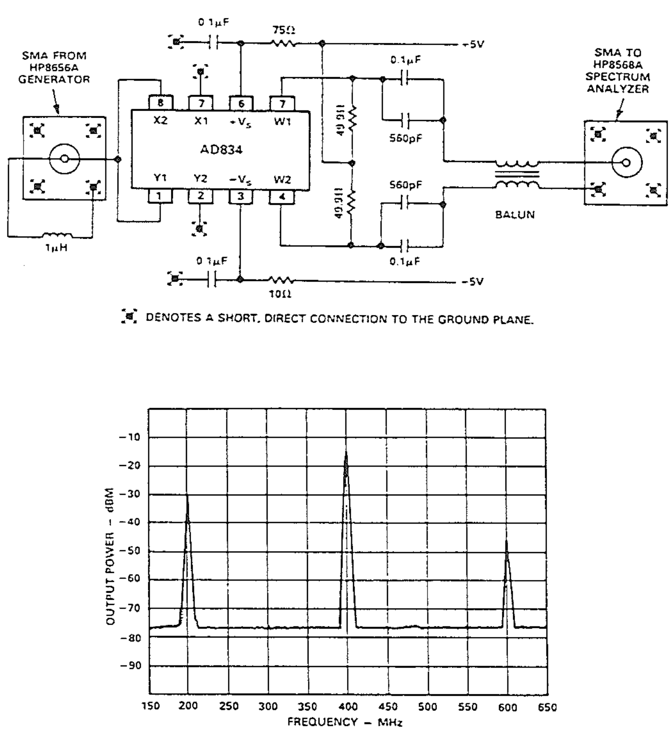
A FUNCTION GENERATOR IN A NEGATIVE-FEEDBACK LOOP GENERATES THE INVERSE FUNCTION

Figure 5.15

If a sinusoidal waveform is applied to a squarer, the output consists of a sinusoidal waveform at twice the frequency, plus a DC level - this necessarily follows from the well-known identity

$$\sin^2\omega t = \frac{(1 - \cos 2\omega t)}{2}$$

This relationship can be used to make a multiplier perform as a frequency doubler as shown in Figure 5.16. If a signal is applied to both ports of an AD834 and the output is AC coupled, the DC component will be lost, and only the component at twice the frequency will appear.

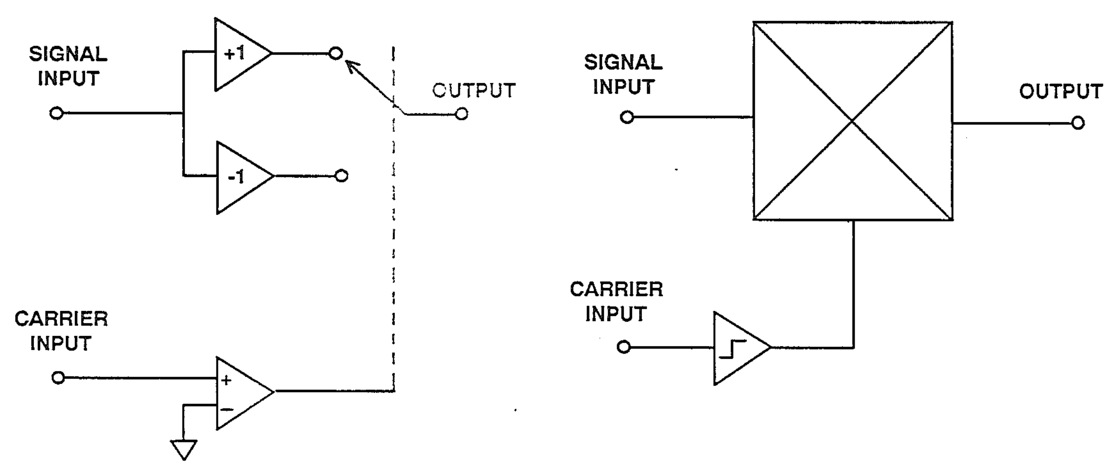


AD834 FREQUENCY DOUBLER

Figure 5.16

MODULATORS

Modulators are closely related to multipliers. The output of a multiplier is the instantaneous product of its inputs - the output of a modulator is the instantaneous product of a signal on one of its inputs (known as the signal input) and the sign of the signal on the other (known as the carrier input).



TWO MODULATOR MODELS

Figure 5.17

A modulator may be modelled as an amplifier whose gain is switched positive and negative by the output of a comparator on its carrier input - or as a multiplier with a high-gain limiting amplifier between the carrier input and one of its ports as shown in Figure 5.17. Both architectures have been used to produce modulators, but the switched amplifier version, although potentially very accurate, tends to be rather slow. Most high speed integrated circuit modulators consist of a Gilbert Cell multiplier with a limiting amplifier in the carrier path.

If two periodic waveforms $A_m \cdot \cos(\omega_m t)$ and $A_c \cdot \cos(\omega_c t)$ are applied to the inputs of a multiplier (with a scale factor of 1 V for simplicity of analysis) then the output will be

$$A_m A_c \cdot [\cos(\omega_m + \omega_c)t + \cos(\omega_m - \omega_c)t].$$

This signal contains signals at the sum and difference frequencies, but not at the original frequencies. (Note that using the cosine formulae rather than sine formulae makes the equations easier to manipulate because $\cos(a) = \cos(-a)$ [which makes sign unimportant during simplification] and because $\cos(0) = 1$, so that for DC signals [when $a\omega t = 0$], $\cos(a\omega t)$ is equal to unity.)

When we say that the original frequencies are not present in the output of a modulator we make the assumption that the modulator is perfectly balanced - i.e. neither its signal port nor its carrier port has any offset. In practice, both ports will have some offset, and so there will be some signal and carrier leak. Trimming offset on the inputs of a modulator will reduce these leaks, but there will always be untrimmable residual leaks which are due to coupling by stray capacitance, and to nonlinearities in the core, rather than to offsets.

CAUSES OF SIGNAL & CARRIER LEAK IN MODULATORS

- OFFSET ON THE SIGNAL PORT CAUSES CARRIER LEAK
- OFFSET ON THE CARRIER PORT CAUSES SIGNAL LEAK
- EVEN WHEN ALL OFFSETS HAVE BEEN TRIMMED OUT THERE IS RESIDUAL SIGNAL AND CARRIER LEAK CAUSED BY STRAY CAPACITANCE AND CORE NONLINEARITIES

Figure 5. 18

This "sum and difference mixer" is the function which we expect of modulators. However, if we use a linear multiplier as a modulator, we find that any noise or modulation on the carrier input appears in the output signal. If we replace a simple multiplier with a modulator, any amplitude variation on the carrier input disappears.

If a signal $A_c \bullet \cos(\omega_c t)$ is clipped in a comparator or limiting amplifier the square wave output which we obtain has the form represented by the Fourier series $K[\cos(\omega_c t) - 1/3\cos(3\omega_c t) + 1/5\cos(5\omega_c t) - 1/7\cos(7\omega_c t) + \ldots]$.

The sum of the series 1 - 1/3 + 1/5 - 1/7..... is $\pi/4$. Therefore the value of K, such that a balanced modulator acts as a unity gain amplifier when a positive DC signal is applied to its carrier input, is $4/\pi$.

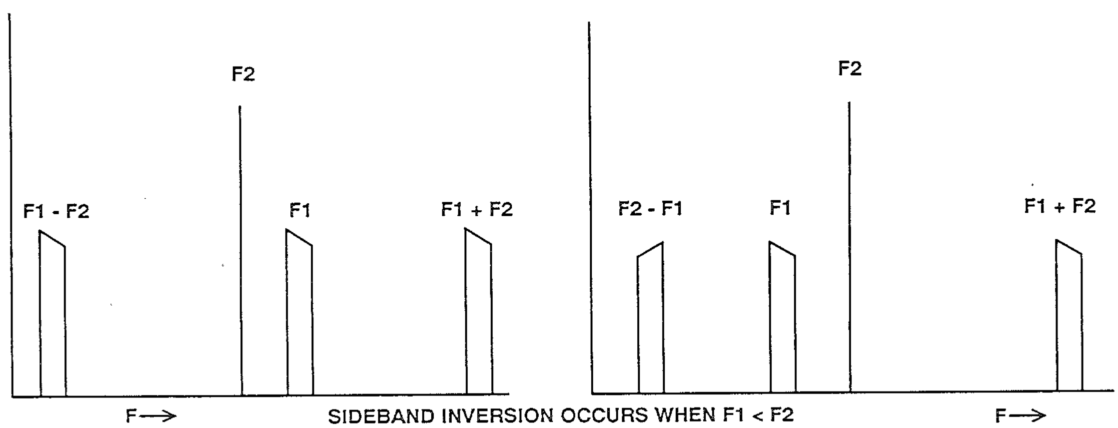
Therefore if a modulator is driven by a signal $A_m \bullet \cos(\omega_m t)$ and a carrier $\cos(\omega_c t)$ (the carrier amplitude is unimportant provided it is great enough to drive the limiting amplifier), then the output will be the product of the signal and the squared carrier above.

This is

$$\begin{aligned}2A_m/\pi[&\cos(\omega_m + \omega_c)t + \cos(\omega_m - \omega_c)t \\ &- 1/3\{\cos(\omega_m + 3\omega_c)t + \cos(\omega_m - 3\omega_c)t\} \\ &+ 1/5\{\cos(\omega_m + 5\omega_c)t + \cos(\omega_m - 5\omega_c)t\} \\ &- 1/7\{\cos(\omega_m + 7\omega_c)t + \cos(\omega_m - 7\omega_c)t\} + \ldots.]\end{aligned}$$

This output contains sum and difference frequencies of the signal and carrier, and of the signal and each of the odd harmonics of the carrier (in the ideal, perfectly balanced modulator, products of even harmonics are not present - in real modulators, which have residual offsets on their carrier ports, low-level even harmonic products are also present, just how low their level depends on the size of the offset). In most applications a filter is used to remove the products of the higher harmonics so that, effectively, the modulator does behave like a multiplier. (In analyzing the above expressions we must remember that cos(A) = cos(-A), so that $\cos(\omega_m - N\omega_c)t = \cos(N\omega_c - \omega_m)t$ so we do not have to worry about "negative frequencies".)

The most obvious application of a modulator is a mixer or frequency changer. If we apply an input signal at F1 and a carrier at F2 to a modulator we find that the output contains signals at the sum and difference frequencies as shown in Figure 5.19. This applies even if the signal is a modulated signal containing a number of components.



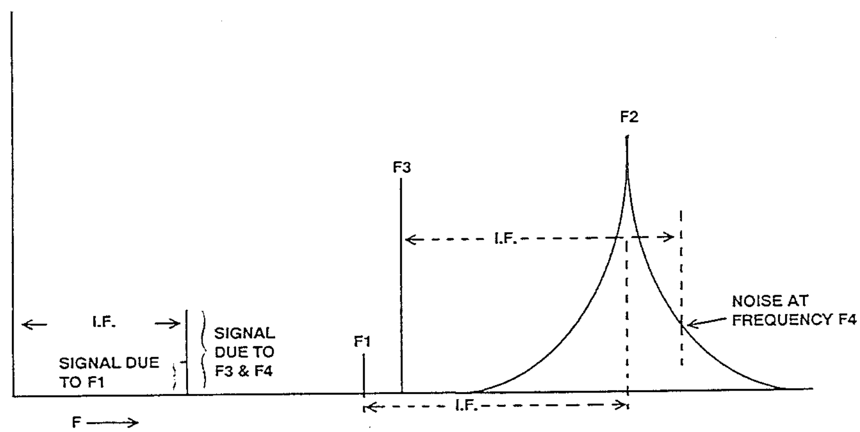
THE MODULATOR AS A MIXER

Figure 5.19

As we have mentioned above, we cannot have "negative frequencies" and so if F1-F2 is negative, what we actually see is a frequency of F2-F1. If F1 is a complex signal containing a number of components, however, we find that if the carrier frequency, F2, is less than F1 then the sidebands are uninverted in both the sum and difference products, but if F2 is greater than F1, then in the difference product, the sidebands are inverted, as shown in the diagram.

The mixer or frequency changer is a key component in most radio receivers. While it is inappropriate to go into a detailed discussion of receiver design in a seminar of this type it is perhaps useful to point out two important features of modulators for use in receivers. These are noise and strong signal performance.[4, 5]

It is self-evident that the mixer used as the front-end mixer of a radio receiver must have low noise, since the signals which it is handling may be less than a microvolt in amplitude, and any significant in-band noise would swamp them. It is less evident that noise in the carrier channel is also important.

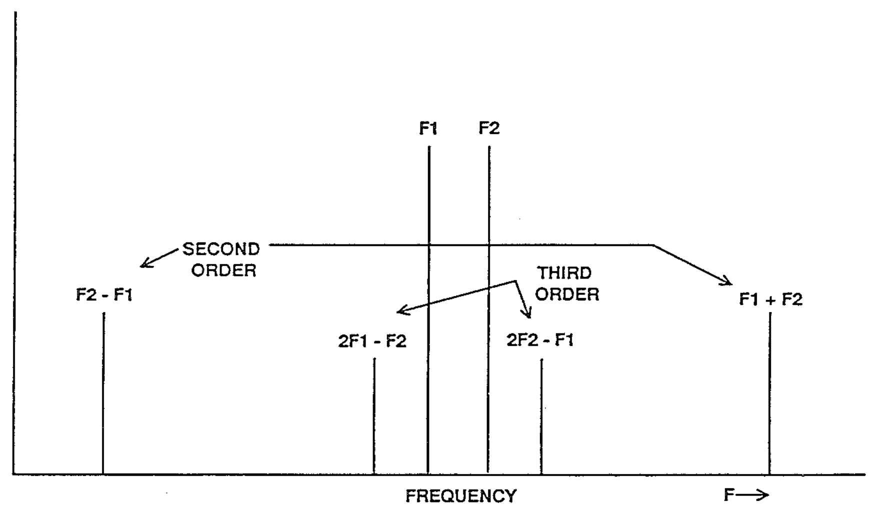


RECIPROCAL MIXING

Figure 5.20

Suppose that we have a mixer with a noisy carrier channel which causes the carrier frequency, F2, to spread out on either side of its center as shown in Figure 5.20. If we are receiving a small wanted signal, F1, then we shall see a small IF output from the mixer at F2-F1. If, however, there is a strong unwanted signal at F3 then the product, F4-F3, of F3 and that part of the broadband carrier noise indicated by F4 on the diagram, will also fall at the IF and, being larger than the wanted IF component, will swamp it. This phenomenon is known as "reciprocal mixing" and can cause a severe limitation on the dynamic range of a receiver. While it is probably more commonly caused by noise or spurious synthesizer sidebands in the oscillator driving the modulator carrier port, it is by no means unknown for it to be caused by noise in the modulator itself, and it should certainly be considered when choosing a mixer for a radio receiver.[6]

In the past, the sensitivity of a radio receiver has been one of its most important features. Today, while sensitivity is still important, the behaviour of the receiver in the presence of strong signals is equally important. The characteristic chosen as a measure of a mixer's performance in this respect is its "third order intermodulation" performance. The key specification is the "Third Order Intercept Point".



INTERMODULATION PRODUCTS
Figure 5.21

Consider a non-linear amplifier with two large input signals, at F1 and F2 as shown in Figure 5.21. The non-linearity gives rise to additional output components at F1+F2 and F1-F2: these are known as "Second Order Intermodulation Products". These second order products mix with the original signals and produce "Third Order Intermodulation Products" at frequencies of 2F1-F2 and 2F2-F1.

These third order intermodulation products are a major nuisance in radio reception, especially in channelized systems, because they fall close to the signals causing them. As an example consider a receiver monitoring a frequency of 145.500 MHz. In Europe this frequency is the calling frequency of the 2 meter Amateur Band. Working channels in this band are separated by 25 KHz. Suppose that close to the receiver monitoring 145.500 MHz are two transmitters, working at 145.400 and 145.450 MHz respectively. The third order intermodulation products of these two frequencies fall at 145.350 and 145.500 MHz. If the receiver is liable to third order intermodulation distortion (IMD) it will respond to the third order products - which it itself produces - and appear to be receiving a signal at 145.500 MHz.

It is impossible to design an amplifier or mixer which is unaffected by third order intermodulation. All that can be done is to minimize the problem. The third order intercept point mentioned above is the parameter which measures how susceptible a device is to third order intermodulation.

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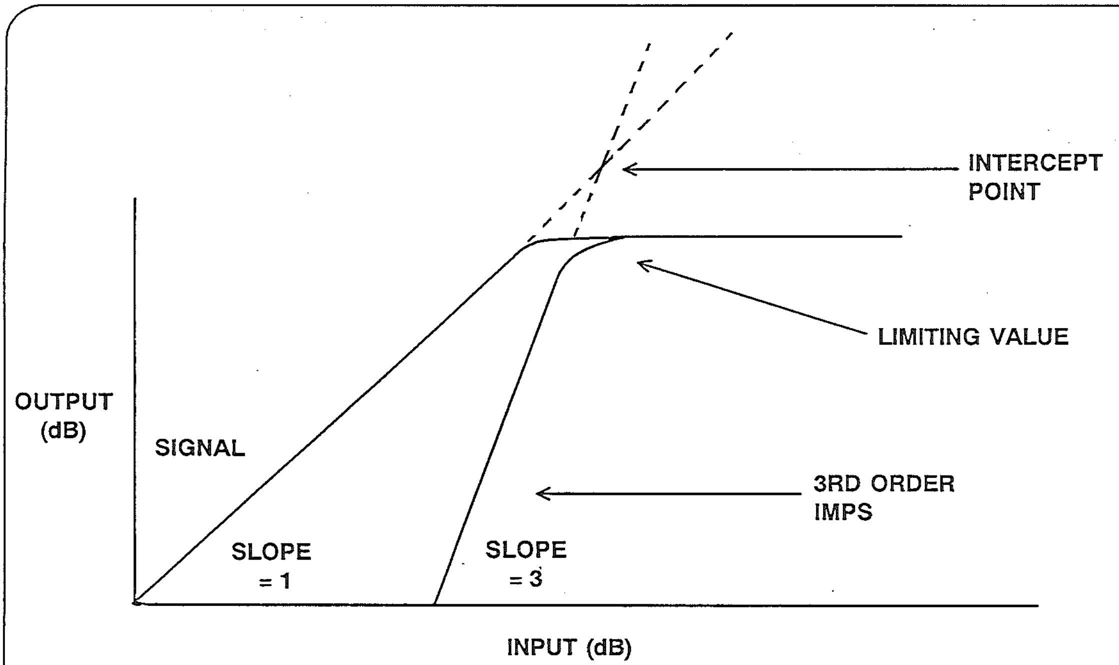


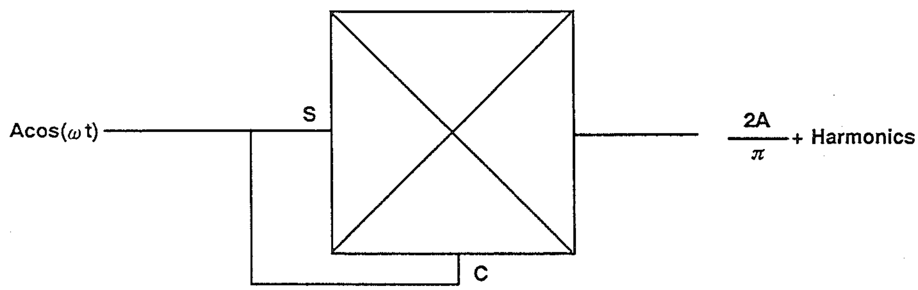
THIRD ORDER INTERCEPT POINT

Figure 5.22

If we plot the input vs the output amplitudes of an amplifier on a log/log (dB/dB) scale as shown in Figure 5.22, we obtain a straight line with a slope of unity. At a certain input level the device saturates ,and the output ceases to rise. If we plot the level of the third order intermodulation products in the output against the level of a two-tone input on the same axes, we obtain a straight line with a slope of 3. This line also ceases to rise when it reaches some limit. If, however, we extend the two straight lines past their limiting values they will eventually cross. The value of the two-tone input at this "intercept point" is the "Third Order Intercept Point" of the device or system being measured.

In mixers for receivers, values of third order intercept point can vary from -15 dBm to over +45 dBm. Any value below 0 dBm is generally considered poor, and good performance requires values of at least +15 dBm and, preferably, more. (For more detailed discussion of intermodulation distortion (IMD) see the Op-Amp section of this seminar.)

The basic mixing property of modulators is also used for many operations where dynamic range is far less important. These include frequency synthesis by mixing, frequency changing with fixed level signals, and sideband generation. Detailed discussion of these operations is beyond the scope of this seminar, but there is one more application which should be considered - precision rectifiers.

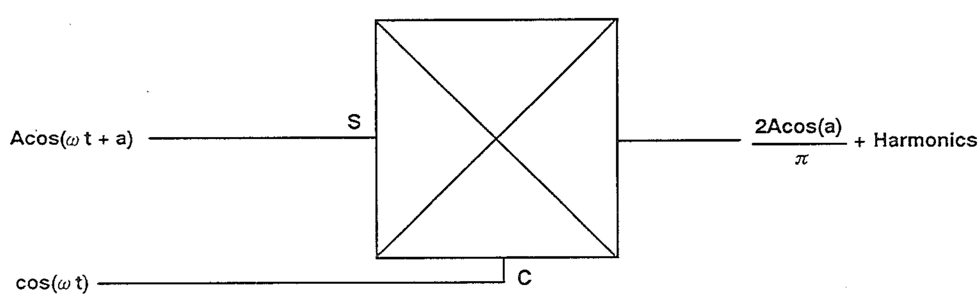


A MODULATOR AS A PRECISION RECTIFIER

Figure 5.23

If an AC signal is applied to both inputs of a modulator as shown in Figure 5.23, the instantaneous output will be equal to the input, if the input is positive, and to the inverse of the input (and therefore still positive), if the input is negative. This arrangement, therefore, behaves as a precision rectifier.

If, instead of applying a signal to both ports of a modulator, a signal is applied to the signal port and a reference signal at the same frequency (but not necessarily the same phase) to the carrier port, then the output will be proportional both to the amplitude of the signal input and the cosine of their phase difference. In this mode a modulator acts as a phase-sensitive rectifier. (See Figure 5.24).



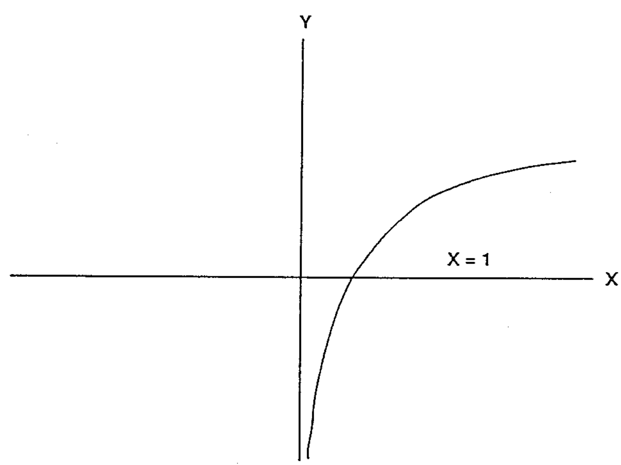
A MODULATOR AS A PHASE-SENSITIVE RECTIFIER

Figure 5.24

LOGARITHMIC AMPLIFIERS

The term "Logarithmic Amplifier" (generally abbreviated to "log amp" - which term will be used in this seminar) is something of a misnomer, and "Logarithmic Converter" would be a better description. The conversion of a signal to its equivalent logarithmic value involves a nonlinear operation, the consequences of which can be confusing if not fully understood. It is important to realize that many of the familiar concepts of linear circuits are irrelevant to log amps. For example the incremental gain of an ideal log amp approaches infinity as the input tends to zero, and a change of offset at the output of a log amp is equivalent to a change of amplitude at its input - not a change of input offset.

For the purposes of simplicity in our initial discussions we shall assume that both the input and the output of our log amp are voltages, although there is no particular reason why logarithmic current, transimpedance, or transconductance amplifiers should not also be designed.



GRAPH OF Y = LOG(X)

Figure 5.25

If we consider the equation $y = \log(x)$ we find that every time x is multiplied by a constant A, y increases by another constant A1. Thus if $\log(K) = K1$ then $\log(AK) = K1+A1$, $\log(A^2K) = K1+2A1$, and $\log(K/A) = K1-A1$, etc. This gives a graph as shown in Figure 5. 25, where y is zero when x is unity, y approaches minus infinity as x approaches zero, and which has no values of x for which y is negative.

On the whole log amps do not behave in this way. Apart from the difficulties of arranging infinite negative output voltages such a device would not, in fact, be very useful. A log amp must satisfy a transfer function of the form

$$V_{out} = V_y \log(V_{in}/V_x)$$

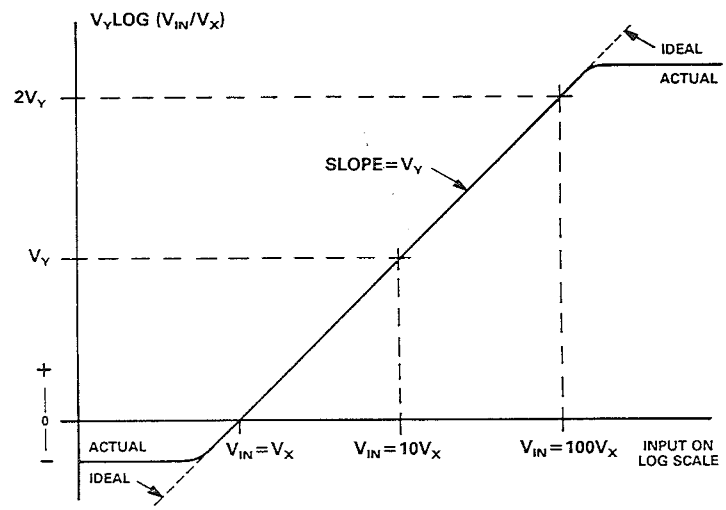
over some range of input values which

may vary from 100:1 (40 dB) to over 1,000,000:1 (120 dB). This ratio is known as the dynamic range of the log amp.

With inputs very close to zero, log amps cease to behave logarithmically, and most then have a linear V_{in}/V_{out} law. This behaviour is often lost in device noise.

Noise often provides the limiting factor to the dynamic range of a log amp.

The constant V_y has the dimensions of voltage because the output is a voltage. The input, V_{in}, is divided by a voltage, V_x, because the argument of a logarithm must be a simple ratio.



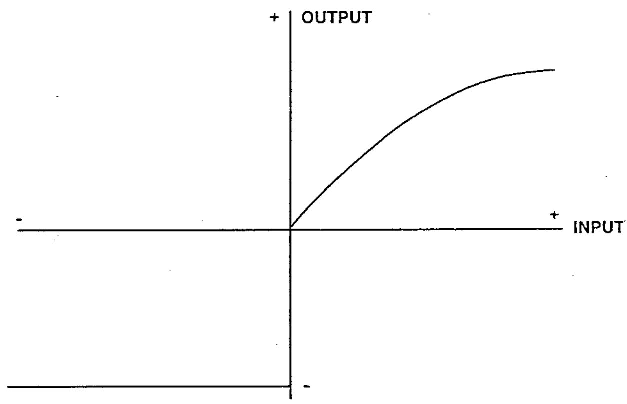
LOG AMP TRANSFER FUNCTION

Figure 5.26

A graph of the transfer characteristic of a log amp is shown in Figure 5.26. The scale of the horizontal axis (the input) is logarithmic, and the ideal transfer characteristic is a straight line. When $V_{in} = V_x$ the logarithm is zero ($\log 1 = 0$). V_x is therefore known as the "intercept voltage" of the log amp because the graph crosses the horizontal axis at this value of V_{in}.

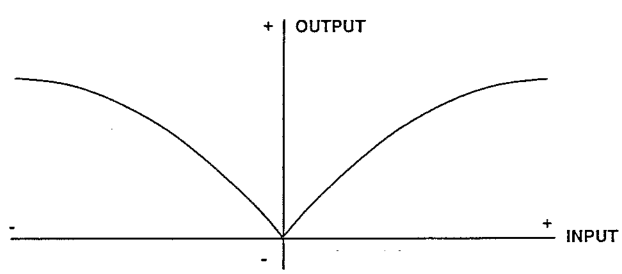
The slope of the line is proportional to V_y. When setting scales we use logarithms to the base 10 because this simplifies the relationship to decibel values: when $V_{in} = 10\,V_x$ the logarithm has the value of 1, so the output voltage is V_y. When $V_{in} = 100\,V_x$ the output is $2\,V_y$ and so forth. V_y can therefore be viewed either as the "slope voltage" or as the "volts per decade factor".

The logarithm function is indeterminate for negative values of x. Log amps can respond to negative inputs in three different ways:



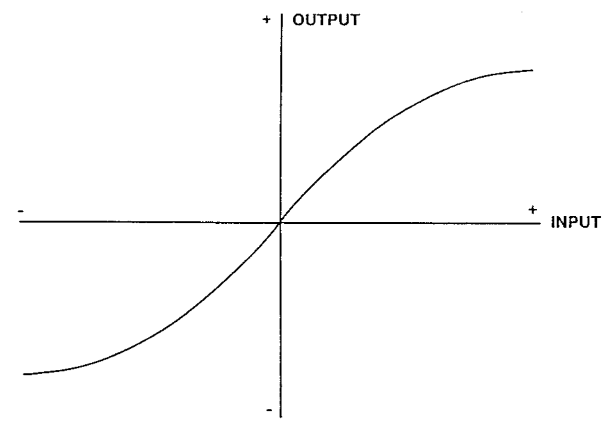
BASIC LOG AMP
(SATURATES NEGATIVE WITH NEGATIVE INPUT)
Figure 5.27

[1] They can malfunction or give a full-scale negative output as shown in Figure 5.27. This is the simplest response and the closest match to the mathematical model - but it is not necessarily the most useful. Such a log amp might be described as a basic log amp.



DETECTING LOG AMP
(OUTPUT POLARITY INDEPENDENT OF INPUT POLARITY)
Figure 5.28

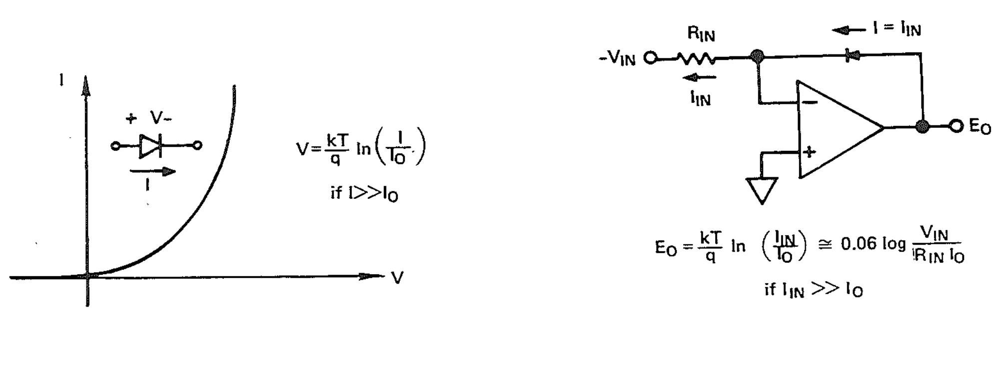
[2] They can give an output which is proportional to the log of the absolute value of the input and disregards its sign as shown in Figure 5.28. This type of log amp can be considered to be a full-wave detector with a logarithmic characteristic, and is often referred to as a detecting log amp.



LOG VIDEO AMP OR "TRUE LOG AMP"
(SYMMETRICAL RESPONSE TO POSITIVE OR NEGATIVE SIGNALS)
Figure 5.29

[3] They can give an output which is proportional to the log of the absolute value of the input and has the same sign as the input as shown in Figure 5.29. This type of log amp can be considered to be a video amp with a logarithmic characteristic, and may be known as a logarithmic video amplifier or, sometimes, a "true log amp".

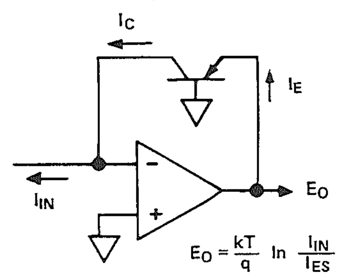
There are three basic architectures which may be used to produce log amps: the basic diode log amp, the successive detection log amp, and the "true log amp" which is based on cascaded semi-limiting amplifiers.



THE DIODE/OP-AMP LOG AMP

Figure 5.30

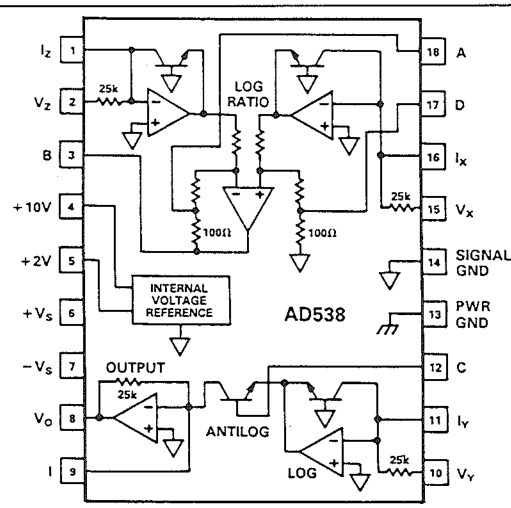
The voltage across a silicon diode is proportional to the logarithm of the current through it. If a diode is placed in the feedback path of an inverting op-amp, the output voltage will be proportional to the log of the input current. (See Figure 5.30)



TRANSISTOR LOG AMP

Figure 5.31

In practice, a combination of bulk resistivity (at high currents) and surface leakage (at low currents) limits the dynamic range of a diode log amp of this type to 40-60 dB. Fortunately, replacing the diode with a diode-connected transistor can easily increase the dynamic range to 120 dB or more. (See Figure 5.31.)

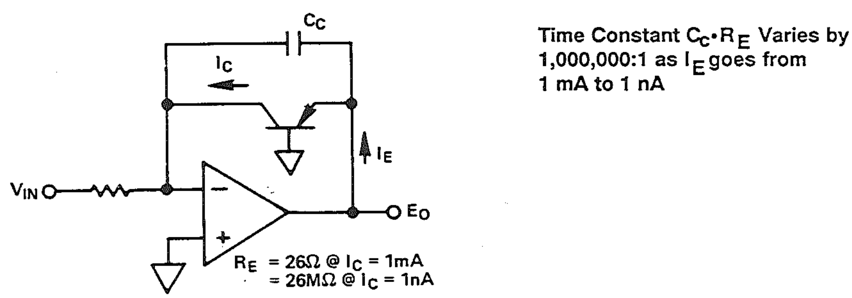


$$V_{out} = V_Y \left(\frac{V_Z}{V_X}\right)^m$$

THE AD538

Figure 5.32

This type of log amp has three disadvantages: (1) both the slope and intercept are temperature dependent; (2) it will only handle unipolar signals; and (3) its bandwidth is both limited and dependent on signal amplitude. Where several such log amps are used on a single chip to produce an analog computer which performs both log and antilog operations, the temperature variation in the log operations is unimportant, since it is compensated by similar variation in the antilogging. This makes possible the AD538, a monolithic analog computer which can multiply, divide and raise to powers (see Figure 32). Where actual logging is required, however, the AD538 and similar circuits require temperature compensation.[7]



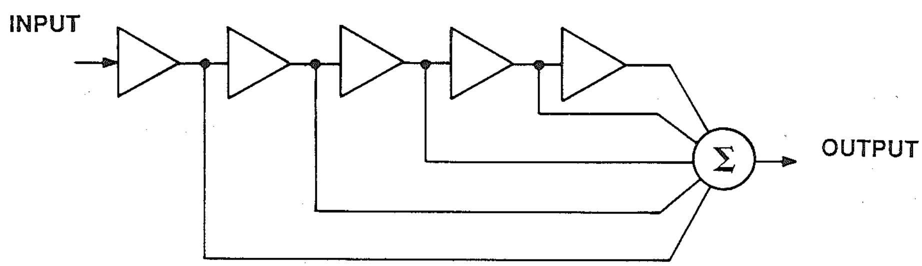
LOG AMP BANDWIDTH IS AMPLITUDE DEPENDENT

Figure 5.33

The major disadvantage of this type of log amp for high frequency applications, though, is its limited frequency response - which cannot be overcome. However carefully the amplifier is designed, there will always be a residual feedback capacitance, C_c (often known as "Miller" capacitance[8]), from output to input, which limits the high frequency response. (See Figure 5.33.)

What makes this Miller capacitance particularly troublesome is that the impedance of the transistor emitter-base junction is inversely proportional to the current flowing in it - so that if the log amp has a dynamic range of 1,000,000:1, then its bandwidth will also vary by 1,000,000:1. In practice, the variation is less because other considerations limit the large signal bandwidth, but it is very difficult to make a log amp of this type with a bandwidth, at small signals, greater than a few hundred KHz.

5



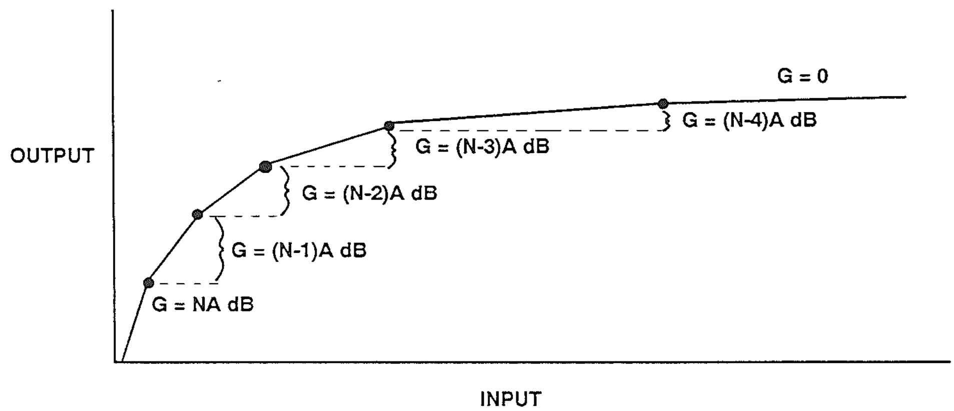
BASIC MULTI-STAGE LOG AMP ARCHITECTURE

Figure 5.34

For high frequency applications, therefore, detecting and "true log" architectures are used. Although these differ in detail, the general principle behind their design is common to both: instead of one amplifier having a logarithmic characteristic, these designs use a number of similar cascaded linear stages having well-defined large signal behaviour.

Consider N cascaded limiting amplifiers, the output of each driving a summing circuit as well as the next stage as shown in Figure 5.34. If each amplifier has a gain of A dB the small signal gain of the strip is NA db. If the input signal is small enough for the last stage not to limit the output of the summing amplifier will be dominated by the output of the last stage.

As the input signal increases the last stage will limit. It will now make a fixed contribution to the output of the summing amplifier, but the incremental gain to the summing amplifier will drop to (N-1)A db. As the input continues to increase, this stage in turn will limit and make a fixed contribution to the output, and the incremental gain will drop to (N-2)A db, and so forth - until the first stage limits, and the output ceases to change with increasing signal input.



BASIC MULTI-STAGE LOG AMP RESPONSE (UNIPOLAR CASE)

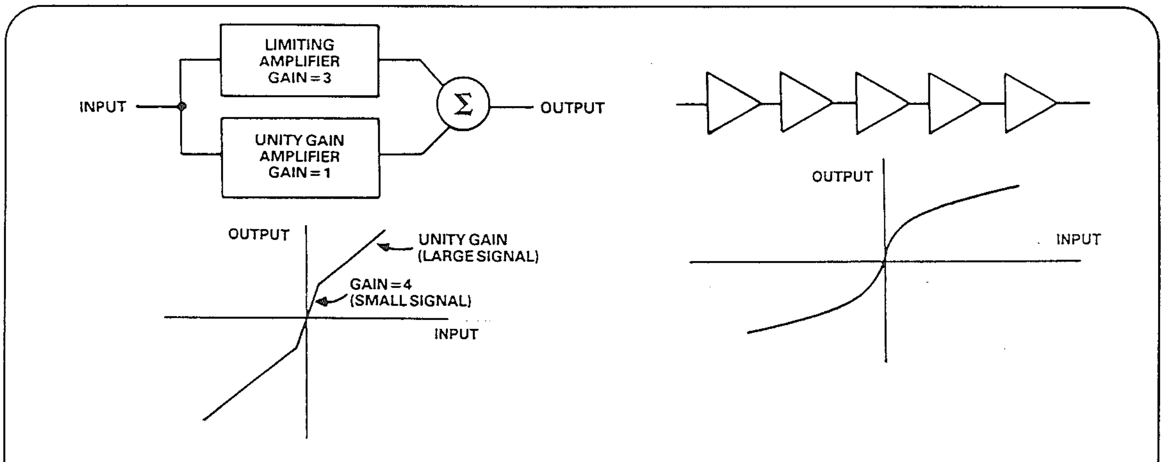
Figure 5.35

The response "curve" is thus a set of straight lines as shown in Figure 5.35. The total of these lines, though, is a very good approximation to a logarithmic curve, and in practical cases is an even better one because few limiting amplifiers, especially high frequency ones, limit quite as abruptly as this model assumes.

The choice of gain, A, will also affect the log linearity. If the gain is too high the log approximation will be poor, if it is too low too many stages will be required. Generally gains of 10-12 dB (3x to 4x) are chosen.

This is, of course, an ideal and very general model - it demonstrates the principle but its practical implementation at very high frequencies is difficult. Assume that there is a delay in each limiting amplifier of t ns (this delay may also change when the amplifier limits but let's consider first order effects!). The signal which passes through all N stages will undergo delay of Nt ns, while the signal which only passes one stage will be delayed only t ns. This means that a small signal is delayed by Nt ns while a large one is "smeared", and arrives spread over Nt ns. A nanosecond equals a foot at the speed of light, so such an effect represents a spread in position of Nt feet in the resolution of a radar system - which may be unacceptable in some systems (for most log amp applications this is not a problem).

A solution is to insert delays in the signal paths to the summing amplifier, but this can become complex. Another solution is to alter the architecture slightly so that instead of limiting gain stages we have stages with small signal gain of A and large signal (incremental) gain of unity (0 dB). We can model such stages as two parallel amplifiers, a limiting one with gain, and a unity gain buffer, which together feed a summing amplifier.

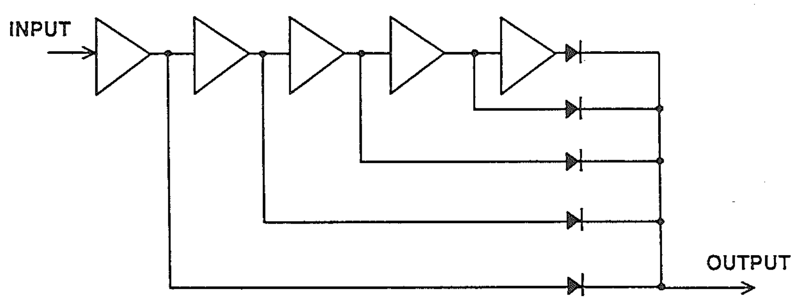


STRUCTURE AND PERFORMANCE OF TRUE LOG AMPLIFIER ELEMENT AND OF A LOG AMP FORMED BY CASCADING SEVERAL SUCH ELEMENTS

Figure 5.36

Figure 5.36 shows that such stages, cascaded, form a log amp without the necessity of summing from individual stages.

Both the multi-stage architectures described above are video log amplifiers or "true log amplifiers", but the commonest type of high frequency log amplifier is the successive detection log amp as shown in Figure 5.37.



Detectors may be full- or half-wave but should be current output devices (not simple diodes) so that the detector outputs may be summed without additional summing components being necessary.

SUCCESSIVE DETECTION LOGARITHMIC AMPLIFIER

Figure 5.37

The successive detection log amp consists of cascaded limiting stages, as in the first example, but instead of summing their outputs directly, these outputs are applied to detectors, and the detector outputs are summed. If the detectors have current outputs, the summing process may involve no more than connecting all the detector outputs together.

Log amps using this architecture have two outputs: the log output and a limiting output. In many applications the limiting output is not used, but in some (FM receivers with "S"- meters, for example), both are necessary.

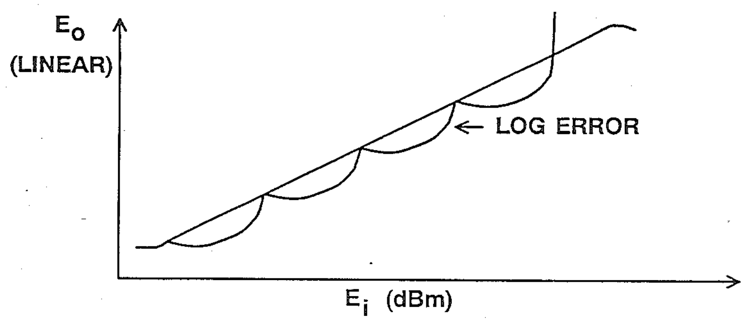
The log output of a successive detection log amplifier generally contains amplitude information, and the phase and frequency information is lost. This is not necessarily the case, however, if a half-wave detector is used, and attention is paid to equalizing the delays from successive detectors - but the design of such log amps is demanding.

The specifications of log amps will, of course, include noise, dynamic range, frequency response (some of the amplifiers used as successive detection log amp stages have low frequency as well as high frequency cutoff), the slope of the transfer characteristic (which is expressed as V/dB or mA/dB depending on whether we are considering a voltage- or current-output device), the intercept point (the input level at which the output voltage or current is zero), and the log linearity. (See Figures 5.38 and 5.39.)

KEY PARAMETERS OF LOG AMPS

- **NOISE** The noise referred to the input (RTI) of the log amp - it may be expressed as a noise figure or as a noise spectral density (voltage, current or both) or as a noise voltage, a noise current, or both
- **DYNAMIC RANGE** Range of signal over which the amplifier behaves in a logarithmic manner (expressed in dB)
- **FREQUENCY RESPONSE** Range of frequencies over which the log amp functions correctly
- **SLOPE** Gradient of transfer characteristic in V/dB or mA/dB
- **INTERCEPT POINT** Value of input signal at which output is zero
- **LOG LINEARITY** Deviation of transfer characteristic (plotted on log/lin axes) from a straight line (expressed in dB)

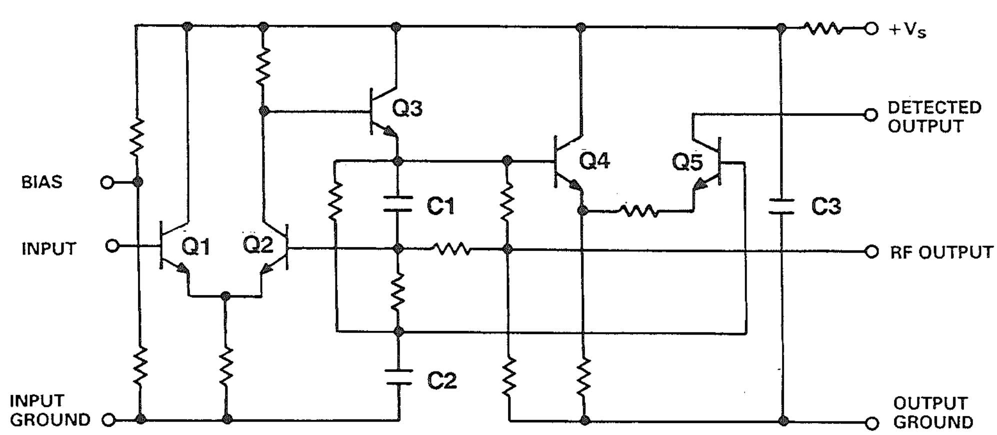
Figure 5. 38



LOG LINEARITY

Figure 5.39

The AD9521 (see Figure 5.40)is a typical high performance successive detection log amplifier component (note that such components are not, themselves, log amplifiers but are components from which log amplifiers may be made). It contains a limiting amplifier which drives both the output and an internal half-wave detector.



AD9521 SIMPLIFIED SCHEMATIC

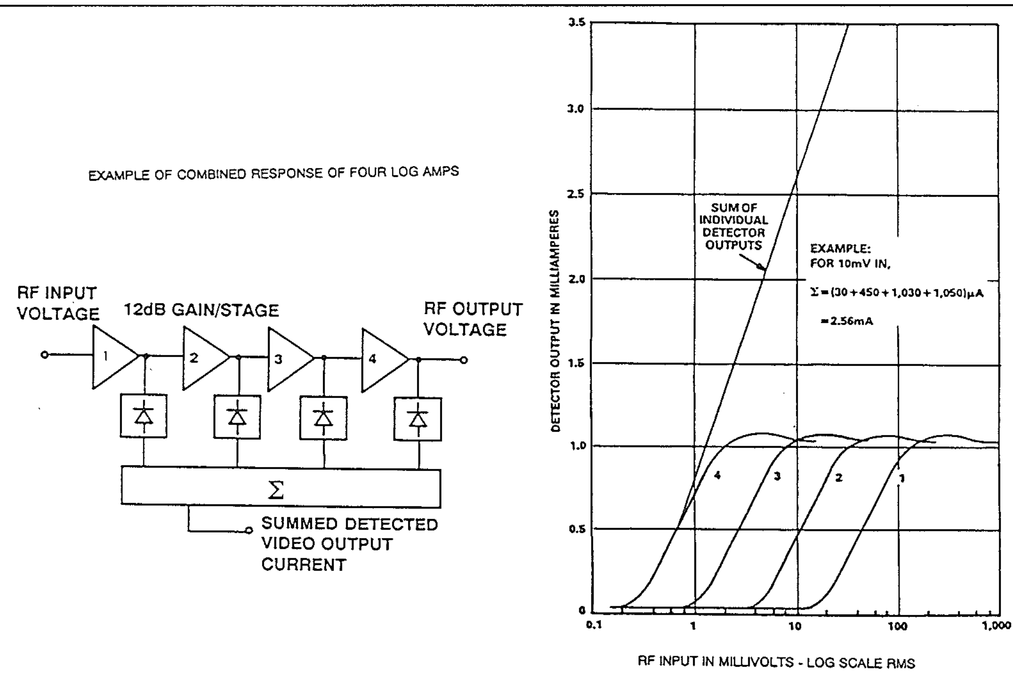
Figure 5.40

AD9521 PERFORMANCE CHARACTERISTICS

- Bandwidth: 7 MHz TO 245 MHz
- Voltage Gain: 12dB
- Maximum Detected Output Current: 0.9 TO 1.1mA at 60 MHz, RF INPUT = 0.5V rms
- Maximum RF Output Voltage: 1.6V P-P
- Maximum Input Before Overload: 1.9V rms
- Power Dissipation: 84mW

Figure 5.41

Figure 5.42 shows the overall transfer function of a log amp made with four cascaded AD9521 devices. The detected current outputs of each stage are plotted against input, as is the sum of all the outputs. It is clear that the sum of these currents approximates to a straight line for inputs between 300 μV and 100 mV - about 48 db.



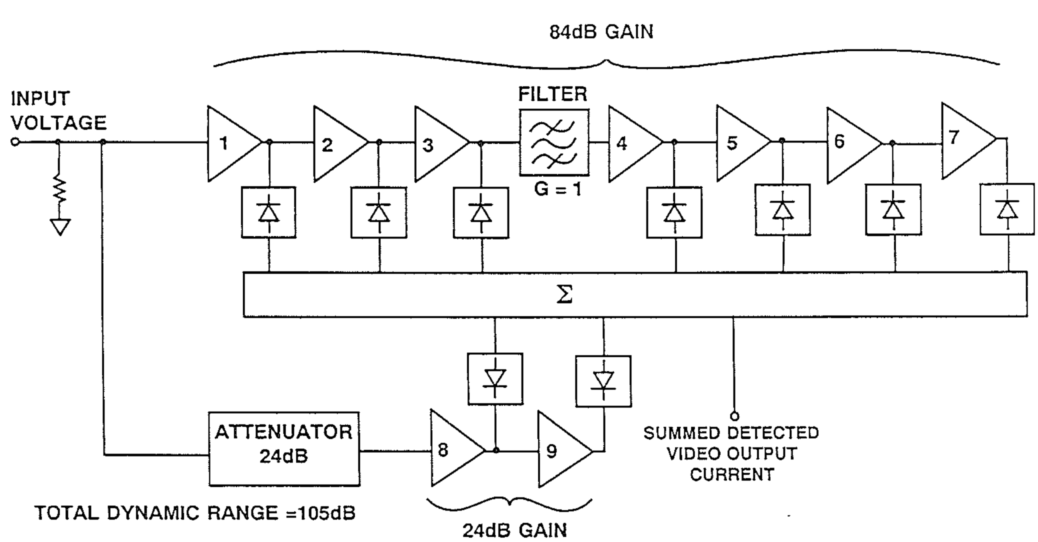
4-STAGE SUCCESSIVE DETECTION LOG STRIP

Figure 5.42

If we add stages, the dynamic range increases by 12 dB with each stage until the strip limits on the noise of its own input stage. This occurs with six stages if they are simply connected together broadband. If NF = 5 dB at 450 Ohms, this gives 70 μV broadband noise [assuming 220 MHz bandwidth] Since an AD9521 limits with 100 mV drive, there must be a gain of less than 1428 (63 dB) to the input of the last stage. At 12 dB/stage this requires five stages so, with the output stage, we cannot have more than six stages without limiting on noise. This gives a dynamic range of less than 70 dB.

We can increase the dynamic range by placing an interstage filter between the third and fourth stages of the strip to limit the bandwidth. If we reduce the bandwidth to 10 MHz, the noise is reduced by the square root of 22 (13.5 dB), so we are still limited to seven stages. The interstage filter used does not affect the accuracy of the log response PROVIDED that it has a voltage gain which is accurately unity throughout its passband.

If we allow for the effects of noise, seven stages will only give us some 80 dB dynamic range. If further dynamic range is required we must use an auxiliary strip. This makes use of the fact that although the output of an AD9521 saturates with 100 mV input, the device operates without problems with inputs of up to 1.9 V.



105 dB LOG AMP USING AN AUXILIARY STRIP

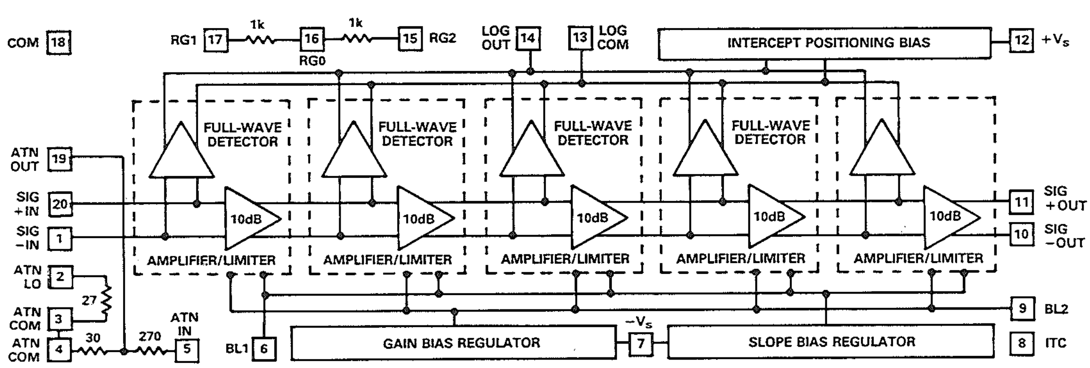
Figure 5.43

If, therefore, we add another two stage strip of AD9521s, with a 24 dB attenuator at its input, in parallel with the existing strip, and summing its outputs to those of the existing strip, we can add another 24 dB of detector range before the input to the main, seven-stage, strip is overloaded. This gives us nine stages for a theoretical dynamic range of 108 dB - in practice it is possible to achieve about 103-105 dB. This log strip configuration is shown in Figure 5.43.

When using AD9521s in long strips there are various considerations of coupling, decoupling, filter design and feedback via the detector pins which must be addressed in any successful design. These are considered in detail in the AD9521 Application Note which is to be published in late 1989.

The AD9521 contains a single stage limiting amplifier/detector. Moreover, it has a low frequency cutoff of about 10 MHz which makes it impossible to use in many lower frequency applications. The AD640 contains five such stages in a single package and its logarithmic performance extends from DC to 145 MHz. Furthermore its amplifier and detector stages are all balanced so that, with reasonably well-considered layout, instability from feedback via supply rails is unlikely. A block diagram of the AD640 is shown in Figure 5.44.

Unlike all previous integrated circuit log amps, the AD640 is laser trimmed to high absolute accuracy of both slope and intercept, and is fully temperature compensated.



BLOCK DIAGRAM OF THE AD640

Figure 5.44

AD640 FEATURES

- 45 dB Dynamic Range - Two AD640s cascadable to 95 dB
- Bandwidth DC to 145 MHz - 120 MHz when cascaded
- Laser-trimmed slope of 1 mA/decade - temperature stable
- Laser-trimmed intercept of 1 mV - temperature stable
- Less than 1 dB log non-linearity
- Balanced circuitry for stability
- Minimal external component requirement

Figure 5.45

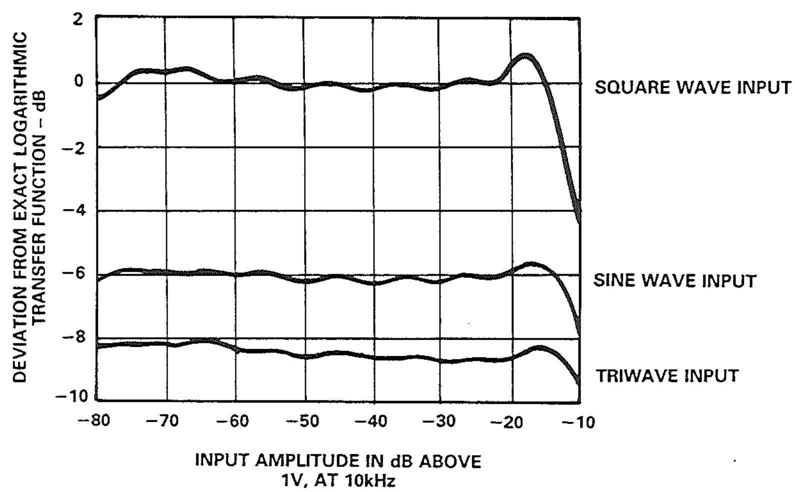
Because of this high accuracy the actual waveform driving the AD640 must be considered when calculating responses. When a waveform passes through a log function generator, the mean value of the resultant waveform changes. This does not affect the slope of the response, but the apparent intercept is modified according to the table. (See Figure 5.46.)

THE EFFECT OF WAVEFORM ON LOG INTERCEPT

Input Waveform	Peak or rms	Intercept Factor	Error (Relative to a dc Input)
Square Wave	Either	1	0.00dB
Sine Wave	Peak	2	-6.02dB
Sine Wave	rms	1.414 ($\sqrt{2}$)	-3.01dB
Triwave	Peak	2.718 (e)	-8.68dB
Triwave	rms	1.569 ($e/\sqrt{3}$)	-3.91dB
Gaussian Noise	rms	1.887	-5.52dB

Figure 5.46

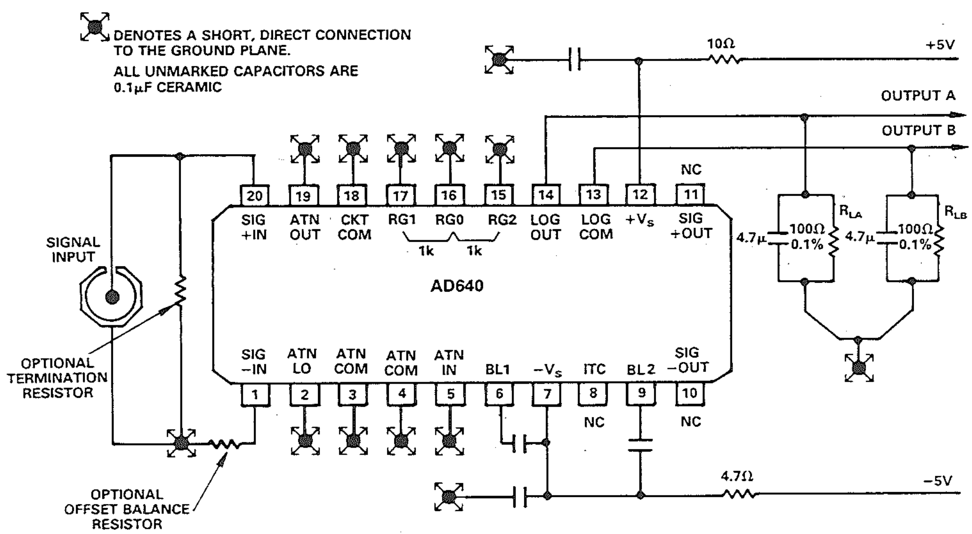
The AD640 is calibrated and laser trimmed to give its defined response to a DC level or a symmetrical 2 KHz square wave, it is also specified to have an intercept of 2 mV for a sine wave input (that is to say a 2 KHz sine wave of amplitude 2 mV pk [NOT pk-pk] gives the same mean output signal as a DC or square wave signal of 1 mV).



THE EFFECT OF WAVEFORM ON LOG LINEARITY AND INTERCEPT

Figure 5.47

The waveform also affects the ripple or non-linearity of the log response. This ripple is greatest for DC or square wave inputs because every value of the input voltage maps to a single location on the transfer function, and thus traces out the full nonlinearities of the log response. By contrast, a general time varying signal has a continuum of values within each cycle of its waveform. The averaged output is thereby "smoothed" because the periodic deviations away from the ideal response, as the waveform "sweeps over" the transfer function, tend to cancel. As is clear in Figure 5.47, this smoothing effect is greatest for a triwave.

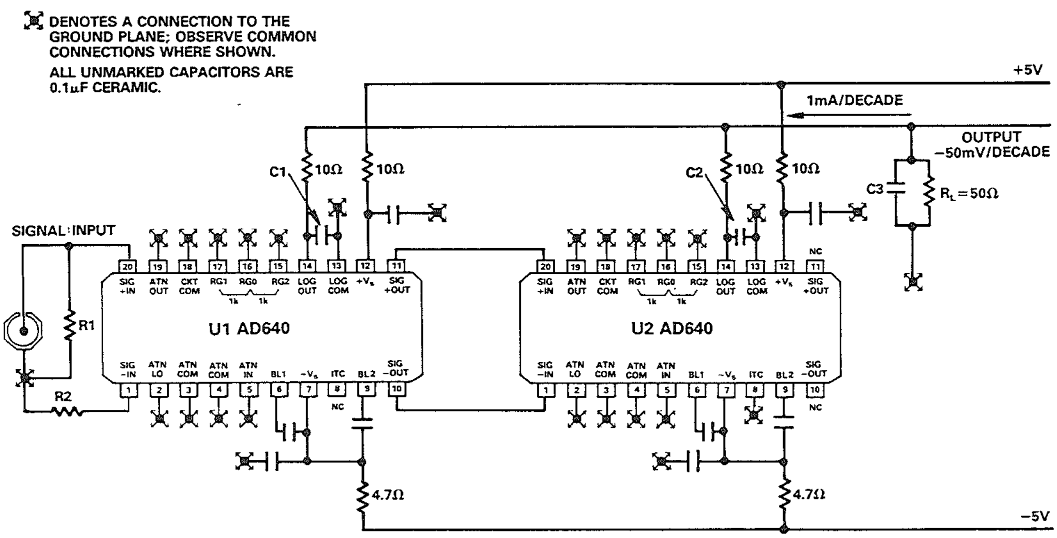


BASIC SINGLE-STAGE AD640 APPLICATION

Figure 5.48

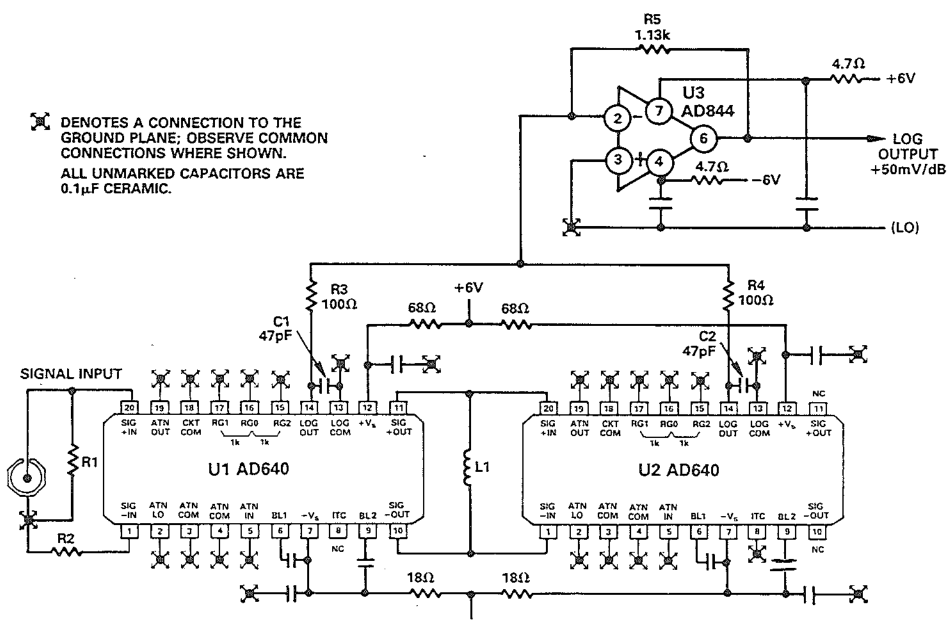
In principle the AD640 is very easy to use (see Figure 5.48) but with such a high wideband gain great care must be taken to keep ground connections short, low inductance and low resistance, to exclude unwanted external signals, and to prevent outputs feeding back to inputs. Detailed applications information is available on the AD640 data sheet.

When two AD640s are cascaded as shown in Figure 5.49, the second will be delivering an output from the noise of the first. If the full potential dynamic range is to be realized the bandwidth must be limited. This may be done with high-pass, low-pass or band-pass filters, depending on the required response, but, as with the AD9521, the voltage gain of these filters in their passband must be unity or there will be a kink in the log response. Figure 5.50 shows a 70 dB log amp for broadband operation from 50 to 150 MHz. The 100 MHz passband limits the possible dynamic range, but the performance is still exceptional. Figure 5.51 shows a 95dB 10Hz to 100KHz log amp using two cascaded AD640s.



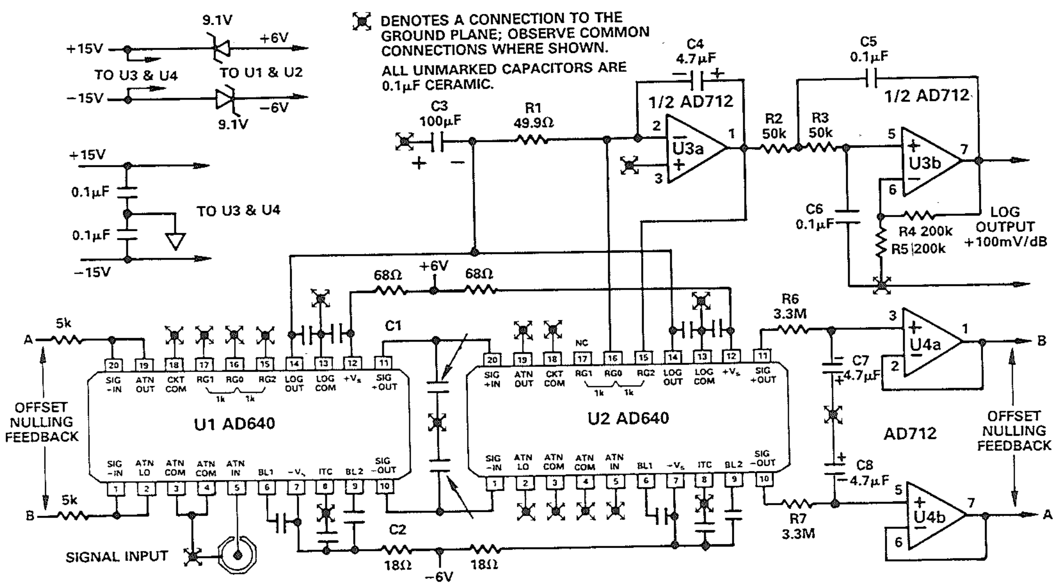
BASIC CASCADED AD640 APPLICATION

Figure 5.49



70 dB LOG AMP FOR 50-150 MHz USING TWO AD640

Figure 5.50



95 dB L.F. LOG AMP (10Hz-100kHZ) USING TWO AD640

Figure 5.51

SPECIFIC REFERENCES

[1]B. Gilbert - **1968 ISSCC Digest of Technical Papers**, pp. 114-115. Feb. 16, 1968.

[2]B. Gilbert - Journal of Solid-State Circuits, Vol. SC-3, Dec. 1968. pp. 353-372.

[3]C. L. Ruthroff, "Some Broadband Transformers" **Proc. I.R.E.** Vol:47, pp.1337-1342 (August 1959).

[4]J. M. Bryant, "Mixers for High Performance Radio" **Wescon 1981: Session 24** (published by Electronic Conventions Inc. Sepulveda Blvd., El Segundo, CA).

[5]P. E. Chadwick, "High Performance I.C. Mixers" **IERE Conference on Radio Receivers & Associated Systems**, Leeds 1981. IERE Conference Publications No:50.

[6]P. E. Chadwick, "Phase Noise, Intermodulation and Dynamic Range" R. F. Technology Expo, Anaheim, January 1986.

[7]Daniel H. Sheingold (ed), "Nonlinear Circuits Handbook" Analog Devices 1974.

[8] John M. Miller, "The Dependence of the Amplification Constant and Internal Plate Circuit Resistance of Three-Electrode Vacuum Tibes upon the Structural Dimensions" Proc. I.R.E. Vol. 8, pp. 64 et. seq, February 1920.

GENERAL REFERENCES

1. Richard Smith Hughes, **Logarithmic Amplifiers**, Artech House, Inc., Dedham, MA, 1986.

2. William L. Barber and Edmund R. Brown, **"A True Logarithmic Amplifier for Radar IF Applications"**, IEEE Journal of Solid State Circuits, Vol. SC-15, No. 3, June 1980, pp. 291-295.

3. **Broadband Amplifier Applications,** Plessey Co. Publication P.S. 1938, Sept. 1984.

4. Tom Tice, "Fast Monolithic Log Amp with Better Than 250MHz Bandwidth", **Analog Dialogue**, Vol. 22-1, 1988, pp. 18-19.

5. M. S. Gay, "SL521 Application Note", Plessey Co., 1966.