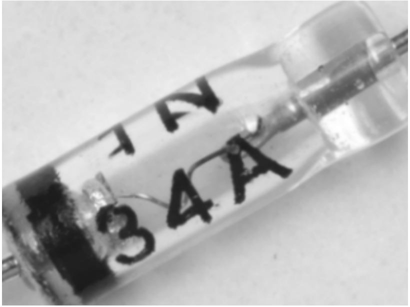
KEVIN'S WEBSURFER HANDBOOK III
FOR CRYSTAL RADIO

DIODES:
SOLID-STATE & VACUUM



Kevin Smith
2011

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http://www.lessmiths.com/~kjsmith/crystal/catalog.shtml

KJ Smith

NOTES:

INTRODUCTION

The contents of this second handbook concern themselves primarily with diodes, diode theory, and aspects of older vacuum diodes. The rectifier diode forms an essential part of the crystal radio circuit, it has a fascinating history, and few parts of the set deliver as much confusion (to me at least) and speculation, comment and even controversy. This document results from my web explorations of diode theory in general and diodes as crystal radio rectifiers in particular. Additionally, I have held a certain fascination with the vacuum diode ever since when reading on the history of radio, I found that this rectifier was actually discovered and patented prior to the lowly galena crystal from which the sets take their name.

In this booklet I have taken the sections and pages that I spend much/most time studying and have arranged them into sequence from basic theory and design considerations to more advanced material and finally on specifically vacuum diodes in particular. Some vacuum tube chapters were especially painful to include as they required pasting in each page of a scanned document. Still, I felt, and feel that this document is especially worthy and useful as a source of vacuum diode theory. Each section is useful and understandable to the hobbyist with moderate to good experience. The handbook will help the beginner to quickly get up to speed and allow the experienced builder to find endless new ideas. This is not a book of Hookups or circuit designs, that is covered in my Catalog of Crystal Hookups, nor is it a tutorial on Crystal Radio, that can be found in my Handbook Volume 1.

Much of the material in this handbook is copyright for which I have not sought permission. Therefore this is not presented for publication or copy, and certainly not for profit. It is only my personal resource. I encourage anyone finding this copy to pursue ON THE WEB the web pages identified within. I include the name of the author and web address of each section. I wish to sincerely thank every author presented for their excellent pages and ask forgiveness for my editing into this handbook.

Kevin Smith
2011

www.lessmiths.com/~kjsmith/crystal/cr0intro.shtml

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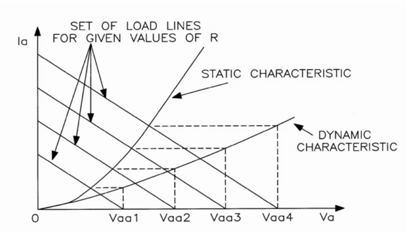
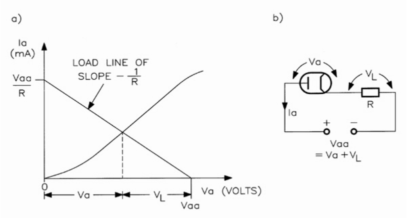
and assuming a constant value for the load Rl then, by drawing a separate load line for each value of supply voltage, the dynamic characteristic can be obtained, as shown below.



Obtaining the dynamic characteristic for a diode valve and its load.

The points on the dynamic characteristic are obtained by projecting, horizontally, the intersection of a load line and the static characteristic until it in turn intersects a vertical line drawn from a supply voltage value. Since the dynamic characteristic is drawn for a range of values of the supply voltage, this implies that the latter is varying, in other words it is an alternating supply rather than DC, as is the case in rectification.

current in the circuit will vary by using a graphical construction.



(a) The load line for a diode valve. (b) the diode in series with a linear resistive load.

The second image shows the diode in series with its load and defines the voltages and currents in question. A 'load line' of slope -l/R is drawn between the two points: Ia = Vaa/R Va = 0 and Ia = 0, Va = Vaa. As in solid state practice, the end points of a load line define zero conduction and maximum conduction, the latter being dependent upon the values of supply voltage and load resistance. Any other points on the load line imply intermediate levels of conduction. By dropping a construction line from the intersection of the load line and the static characteristic, we can see how the total supply voltage Vaa is divided into the two separate voltages Va (the voltage across the diode) and Vl (the voltage across the load).

The Dynamic Characteristic

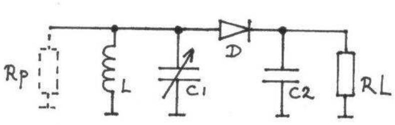
By taking a number of different values of supply voltage Vaa (as would happen if the supply was alternating, for example)

How to build a sensitive crystal receiver

Dick Kleijer crystal-radio.eu
http://www.crystal-radio.eu/engev.htm

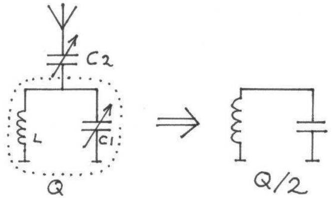
[Calculator included at the bottom of the web page to calculate the components of your receiver for maximum sensitivity at weak signals].

In this article some information about the design of a crystal receiver with maximum sensitivity.



Circuit diagram 1
Circuit diagram of the crystal receiver, which we are going to design for maximum sensitivity at weak signals.
This can be a detector circuit of a 2 circuit receiver.
But also a receiver with loop antenna.

RP represents the losses in coil L and tuner capacitor C1



Circuit diagram 2
If an antenna is connected via a matching network (in this case C2) to the LC circuit, this will reduce the circuit Q.
When optimal matched to the antenna, the Q will be halve the value of the unloaded Q (of L,C1).
For a calculation of the values of C1 and C2, click here.

For the calculation of maximum sensitivity: replace the LC circuit + antenna by a LC circuit with an unloaded Q equal to halve the original value.
Coil value L stays the same.

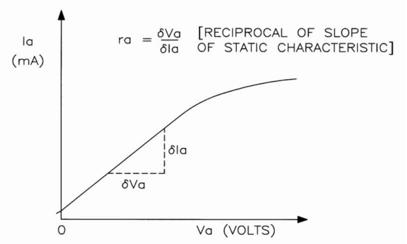
You get maximum sensitivity if there is maximum power transfer from LC detector circuit to load (loudspeaker).
The LC circuit has a certain parallel resistance RP, this is not a real resistor, but a virtual resistor caused by the losses in coil and tuner capacitor.

If we know the Q-factor of the detector circuit, and the induction of the detection coil, we can calculate the value of RP as follows:

RP = 2.pi.f.L.Q (Ohm)

The Anode Slope Resistance ra

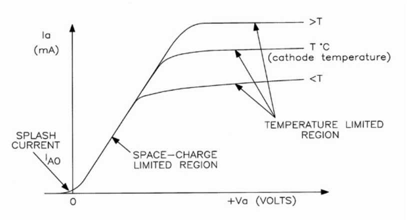
It is worth introducing this parameter at this time since it is one that we shall make use of later in discussing the performance of more complex valves. It is defined as shown below,
Defining anode slope resistance for a diode valve.



and is the value of resistance obtained by dividing a small change in anode voltage by the corresponding change in anode current. It is therefore the reciprocal of the slope of the static characteristic, and varies with the operating point, although fairly constant over much of the space charge limited region. This is a real value of resistance, since it represents the opposition of the valve to alternating quantities.

Series Circuit Operation

It is usual to operate a diode valve, which is clearly a non-linear device, in series with a resistive load, the latter being a linear device. It is possible to predict how the voltages and



The static characteristic of a diode valve.

Above three curves have been drawn for different values of cathode temperature, although in practice, as explained earlier, the cathode is held at a constant temperature.

It is interesting to note that:

A) The current is not exactly zero when the anode voltage is zero, but has a value (Iao) of a few micro-amperes. This is known as the 'splash current' and is the result of a few high energy electrons that manage to cross the inter-electrode gap even without an attracting potential.
B) In the space-charge limited region, the characteristic is nearly linear (actually following the 'three-halves' power law: Ia is proportional to Va-3/2).
C) In the temperature limited region there is little change in la even though there are large changes in Va. This is because the anode is collecting electrons at the same rate as they are being emitted by the cathode.
D) No significant current flows when the anode is negative with respect to the cathode.

pi = 3.14
f = the frequency (Hertz).
L = induction of detector coil (Henry = H).
Q = the quality factor of the unloaded LC circuit.

The value of RP is depending on the frequency, use for instance the value of RP in the middle of the frequency range.
For medium wave for instance, we can use the value of RP at 1 MHz.

Maximum power transfer from LC circuit to load.
If the load resistor RL was directly connected across the LC circuit, it would be simple:
maximum power transfer would occur if RL=RP.
The loaded Q of the LC circuit is then halve the value of the unloaded Q.

But of course we also need a diode between LC circuit and load RL, to demodulate the RF signal.
In the rest of this article we assume the RF signal on the LC circuit is not modulated, so the output of de diode is a DC voltage.

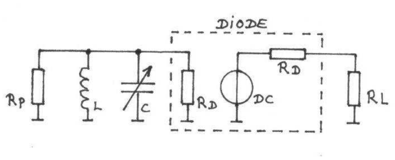
Maximum sensitivity at strong signals.
With a very strong signal across the LC circuit, the diode shall work in the linear detection region.
If the input voltage is high enough, the diode will give only very few power losses, compared to the rectified power.
In the calculation (with strong signals) I assume the diode has no losses at all.
This can in practice not be reached, but it rather simplifies the calculations.

If the diode has no losses, the DC voltage across the load resistor will be equal to the peak voltage of the RF signal.
This peak voltage is 1.41 times the RMS value of the input signal.

In this situation, maximum power transfer to the load RL will occur if:
RL = 2 x RP
The loaded Q of the LC circuit is then halve the value of the unloaded Q.

However with strong signals, maximum sensitivity is not a very important subject, output power is already high enough to be heard.
We can better design the receiver for maximum sensitivity at weak signals.

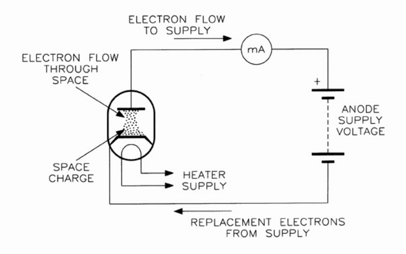
Maximum sensitivity at weak signals.
When receiving (very) weak signals, the diode shall work in the square law detection region.



Circuit diagram of the diode.

With the equivalent circuit of the diode at low signal levels

which may be detected by an ammeter placed in, say, the anode lead. The picture below shows an illustration of this.



The flow of current in a diode valve.

Diode Static Characteristics

We now start getting into the ways in which specifications for thermionic devices are presented. For the diode, these illustrate clearly the dependence of anode current upon anode voltage.

passed forward into semiconductor phraseology - cathode, anode, emitter, collector!).

Since electrons are negative charged particles, they will only be attracted to the anode if this is given a positive potential with respect to the cathode. This explains why the valve only conducts in one direction, from cathode to anode, and not vice versa. It also explains the choice of the word 'valve' to describe the device, since a valve is, by definition, a one-way device. The Americans, however, never cottoned on to this terminology and always refer to them as 'vacuum tubes'.

The magnitude of the current flowing in a diode depends upon the number of electrons emitted and the magnitude of the voltage applied to the anode (known as the anode voltage Va). The amount of electron emission depends upon the temperature of the cathode, which is fixed by the voltage supply to the heater, this being a constant value. The only true variable is, therefore, the anode voltage. The action of the latter in controlling the anode current can be explained as follows.

As we now know from the foregoing, the cathode is normally surrounded by a cloud of electrons known as the space charge. With zero anode voltage there is no current flow, and there is a state of equilibrium between the electrons being emitted and those falling back onto the cathode's surface, The application of a small positive voltage to the anode causes some of the space charge electrons to be attracted to the anode, resulting in a small anode current flow. The gaps created by these electrons leaving the space charge are filled by further emission from the cathode. Electrons arriving at the anode flow to the positive supply terminal, while at the same time an equal number of electrons leave the negative supply terminal for the cathode. This gives rise to a continuous current flow around the circuit,

The input of the diode behaves like a resistor with value RD parallel to the LC circuit.
The diode output is like a DC voltage source in series with a resistor RD.
Maximum power transfer from DC voltage source to load RL occurs when: RL = RD.

In the square law detection region, the detected DC voltage is proportional to square of the RF input voltage.
The power in the load resistor is proportional to the square of the detected DC voltage.
In other words, the power in load RL is proportional to the 4th power of the voltage across the LC circuit.
So it is important to make the voltage across the LC circuit as high as possible, this is done by making the impedance of the loaded LC circuit as high as possible.

If we make all impedances equal, so RL = RD = RP the loaded Q of the LC circuit would be equal to 0.5 times the unloaded Q.
The voltage across the LC circuit is then also 0.5 times the unloaded voltage.
The output voltage is then proportional to: 0.5^4 = 0.0625
If we however make RD and RL 3 times as high, so RL = RD = 3xRP the loaded Q of the LC circuit will be 0.75 times the unloaded Q.
The voltage across the LC circuit is then also 0.75 times the unloaded voltage.
The output power is then proportional to (0.75^4)/3 = 0.3164/3 = 0.1055.
We divide by 3 because the load resistor is now 3 times higher.
So we now have 0.1055 / 0.0625 = 1.6875 times more output power (compared to the situation: RL = RD = RP).
And also, the loaded Q is now 1.5 times higher, so a better selectivity for the receiver.

In the next Excel file: diode.xls the relative output power is calculated for many values of RD and RL, where RD=RL=3xRP is the most sensitive combination.

Conclusion: At weak signals there is maximum power transfer from LC circuit to load, so maximum sensitivity for the receiver if:
RL=RD=3xRP The loaded Q is then 0.75 times the unloaded Q.

Increasing the circuit Q
Increasing the unloaded Q of the LC circuit, will make the impedance of the LC circuit higher.
Because of this the voltage across the LC circuit will increase, the diode then will work more efficient, and output power increases.
More information about increasing the Q, you can find here.
Very good crystal receivers make use of LC circuits with an unloaded Q above 1000.
A high Q also gives a selective receiver.

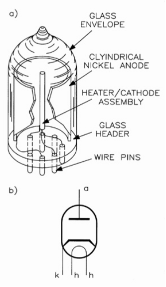
Increasing coil induction
The impedance of the LC circuit can also be increased by increasing the coil induction (μH value).
However making the induction higher, can also result in a lower circuit Q, compensating a part of the impedance increase.
And a lower Q of course gives worse selectivity.
Practical values for induction of medium wave receivers are 200-300 μH.

http://www.r-type.org/articles/art-010b.htm

Valve Technology - A Practical Guide
A series of articles from 1993 by Graham Dixey C.Eng., MIEE republished by kind permission of Maplin Magazine.

The Diode Valve

The diode valve is so called because it has just two electrodes – the cathode and the anode.



(a) Construction of a modern diode valve (indirectly heated type), (b) circuit symbol for a diode valve.

These correspond to the two electrodes of the original diode valve mentioned above, the cathode being the electrode that is heated and emits electrons, and the anode being the electrode that collects the electrons (notice also that these terms have

Choice of the diode

If we know the diode resistance RD, we can calculate the diode "saturation current" (Is value) with the formula:

Is = 0.000086171 x n x TK / RD

Is = saturation current of the diode in A
n = ideality factor of the diode, if you don't know the value, then take n=1.08
TK = temperature in Kelvin (= temperature in ºC + 273)
RD = diode resistance in Ω

Now search for a diode which has a Is value close to the calculated Is value.
Or connect several diodes with a low Is parallel, to reach together the desired value.
More information about diodes you can find here.

Diode connected to a tap on the coil.

The best sensitivity is always reached if the diode is connected to top of the LC circuit, and RL=RD= 3xRP.
But not always it will be possible to make the load resistance RL high enough.
If we don't have a load (audio transformer) with a high enough impedance, the LC circuit will be loaded to heavy, and the Q will reduce.
A method to prevent this, is to connect the diode on a tap somewhere on the coil, instead of the top of the coil.
The Q of the loaded circuit then will increase.
With the calculator at the bottom of this page (click here) the effect on output power can be calculated when using a tap on the coil.

Audio transformer.

When making a receiver with a high Q factor, the load resistance will have to be very high, for instance some Mega-Ohms.
In that case we need a transformer which transforms this high impedance down to the impedance of the used loudspeaker.
Transformers with such a high input impedance are hardly to find.
They can however be self made, look for instance at my transformer unit 2
At high input impedances, in practice both efficiency and bandwidth of a transformer will reduce.

Loudspeaker

Of course we use a loudspeaker with a sensitivity as high as possible.
More information about this, you can find here.

Calculate optimal components for detector circuit

With calculator 1, diode and load resistor are calculated for which the sensitivity of the detector circuit at weak signals is maximum.
Fill in the yellow coloured fields, and click on "calculate calculator 1".

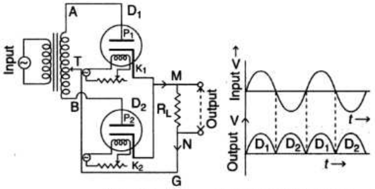
These values can then be taken over in calculator 2.
Then you can self change in calculator 2 all parameters of the detector circuit (all yellow fields), and calculate the effect of this on sensitivity and loaded Q.

2. A rectifier is always followed by electronic circuits called filter circuits which allow only D.C. to pass through them and by pass the A.C. Thus a rectifier-filter combination gives a D.C. output.

Applications of Diode

The two main applications of diode are :

(a) As a rectifier (b) As a detector



(i) Circuit (ii) Input-Output waveforms

positive. Therefore, the second diode conducts but the first remains passive. Thus, the two diodes conduct alternately. But in both half cycles of A.C., the current in load resistance R_L flows in the same direction. So, we get continuous D.C. at output. For full-wave rectifier,

Note :

$$I_{D.C.} = \frac{2I_0}{\pi}$$

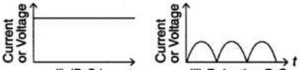
$$I_{rms} = \frac{I_0}{\sqrt{2}}$$

$$\eta_{max} = 81{\cdot}2\%$$

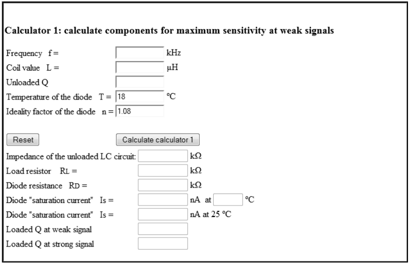
Ripple factor, $r = 0{\cdot}48$

The output of rectifier is fluctuating or pulsating. To make it smooth, filters are used.

Notes : 1. The output obtained from a half wave or full wave rectifier is not a pure D.C. but it is a pulsating D.C. (mixture of A.C. and D.C.)



(i) (D.C.) (ii) Pulsating D.C.



Calculator 1: calculate components for maximum sensitivity at weak signals

Frequency f = kHz
Coil value L = µH
Unloaded Q
Temperature of the diode T = 18 °C
Ideality factor of the diode n = 1.08

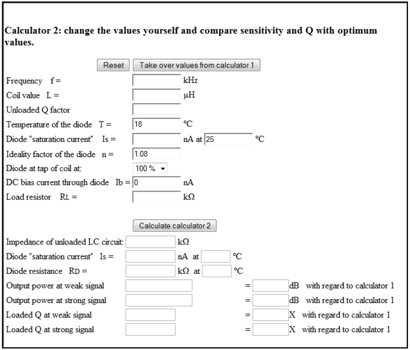
Reset Calculate calculator 1

Impedance of the unloaded LC circuit: kΩ
Load resistor RL = kΩ
Diode resistance RD = kΩ
Diode "saturation current" Is = nA at °C
Diode "saturation current" Is = nA at 25 °C
Loaded Q at weak signal
Loaded Q at strong signal

Calculator 1: calculate components for maximum sensitivity at weak signals

Frequency f = kHz
Coil value L = µH
Unloaded Q
Temperature of the diode T = ºC
Ideality factor of the diode n =

Impedance of the unloaded LC circuit: kΩ
Load resistor RL = kΩ
Diode resistance RD = kΩ
Diode "saturation current" Is = nA at ºC
Diode "saturation current" Is = nA at 25 ºC
Loaded Q at weak signal
Loaded Q at strong signal


Calculator 2: change the values yourself and compare sensitivity and Q with optimum values.

Reset | Take over values from calculator 1

Frequency f = kHz
Coil value L = µH
Unloaded Q factor
Temperature of the diode T = 18 °C
Diode "saturation current" Is = nA at 25 °C
Ideality factor of the diode n = 1.08
Diode at tap of coil at: 100 %
DC bias current through diode Ib = 0 nA
Load resistor RL = kΩ

Calculate calculator 2

Impedance of unloaded LC circuit: kΩ
Diode "saturation current" Is = nA at °C
Diode resistance RD = kΩ at °C
Output power at weak signal = dB with regard to calculator 1
Output power at strong signal = dB with regard to calculator 1
Loaded Q at weak signal = X with regard to calculator 1
Loaded Q at strong signal = X with regard to calculator 1

Calculator 2: change the values yourself and compare sensitivity and Q with optimum values.

Frequency f = kHz
Coil value L = µH
Unloaded Q factor
Temperature of the diode T = ºC
Diode "saturation current" Is = nA at ºC
Ideality factor of the diode n =
Diode at tap of coil at:
DC bias current through diode Ib = nA
Load resistor RL = kΩ

Impedance of unloaded LC circuit: kΩ

Operation

- A.C. input is applied the plate 'P' becomes positive and negative alternately.
- For the positive half cycle of input, the plate is positive with respect of cathode. So that diode conducts and plate current flows through the tube load R_L and secondary. So output occurs across load. R_L as shown in fig. (ii).

Important Points

- During – ve half cycle of A.C. input, →Plate is negative with respect to cathode. So the diode does not conduct, and no voltage appears across output.
- For half wave rectifier; (I_0 is peak value of current)
  - → Average value of current $I_{D.C.} = I_0/\pi$
  - → Root mean square value of current $I_{rms} = I_0/2$
  - → Maximum efficiency $\eta_{max} = 40{\cdot}6\%$
  - → Ripple factor $r = 1{\cdot}21$

(B) **Full wave rectifier**—It converts full A.C. into D.C. In this rectifier two diodes are used which conduct alternately. The output is obtained across the load resistance R_L. Here in first half cycle of A.C., plate of first diode remains positive and plate of second diode remains negative. Hence, the first diode conducts and the second does not. In second half cycle of A.C., plate of the first diode becomes negative and that of second diode becomes

Rectifier

Rectifier is a device which converts an A.C. into a D.C. The process is called **Rectification**. Rectifiers are of two types :

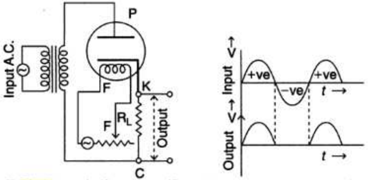
1. **Half wave rectifier**—Which conducts only during + ve half cycle of input A.C.

2. **Full wave rectifier**—Which conducts during full cycle of input A.C.

(A) **Half wave rectifier** (in half wave rectifier only one diode valve is used)

Circuit

- The input A.C. voltage to be rectified is applied to the primary of a transformer.
- The secondary of transformer is connected in series with load R_L between plate and cathode of diode (as shown in fig.).



(i) Diode as a half-wave rectifier (ii) Input and output waveforms

Diode "saturation current" Is = nA at ºC
Diode resistance RD = kΩ at ºC
Output power at weak signal = dB with regard to calculator 1
Output power at strong signal = dB with regard to calculator 1
Loaded Q at weak signal = X with regard to calculator 1
Loaded Q at strong signal = X with regard to calculator 1

Example 1.

We have a LC circuit with a 200 μH coil, the unloaded Q is 1000 and the frequency is 1000 kHz.
As load we use a 100 kΩ audio transformer, because this is the only one we have.
What is the output power at weak signals compared to the optimal load resistor.

Fill in the following values, in calculator 1:
f = 1000 kHz
L = 200 μH
Q = 1000
Click on "calculate calculator 1" (The optimal load resistor and diode are now calculated).
Click on "take over values from calculator 1"

Change load resistor RL in calculator 2 in 100 kΩ.
Click on "calculate calculator 2"
We now see the sensitivity at weak signals is 9.97 dB lower compared to the optimal load.
It's also interesting to see, the Q at strong signals is only 38

Example 2.
We have 2 receivers, both with a 200 μH coil.
One receiver has an unloaded Q of 100, the other receiver has an unloaded Q of 1000, at 1000 kHz.
Both receivers have optimal diode and load resistor.
What is the difference in output power between the 2 receivers?

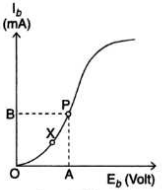
Fill in the following values, in calculator 1:
f = 1000 kHz
L = 200 μH
Q = 1000
Click on "calculate calculator 1" (The optimal load resistor and diode are now calculated).
Click on "take over values from calculator 1"

Reduce the Q in calculator 1 to: 100.
Click on "calculate calculator 1" (The optimal load resistor and diode are now calculated).
Click on "calculate calculator 2".

The result is, the high Q receiver gives 1000 times more output power (+30 dB) at weak signals.
At strong signals, the difference in output power is 10 times (+10 dB).

Example 3.
We have a simple crystal receiver, with a 200 μH coil, and an unloaded Q of 200.
As diode we use a germanium diode with Is = 400 nA (at 25 ºC).
The load resistor is 47 kΩ.

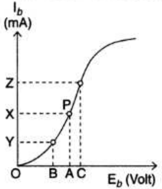
as D.C. plate resistance. The D.C. plate resistance R_p is given by



$$R_p = \frac{OA}{OB}$$

Note : Since plate characteristic curve is not a straight line—Therefore, D.C. plate resistance is variable. Hence, D.C. plate resistance may be calculated at the actual operating point.

(ii) **A. C. plate resistance**—The ratio of a small change in plate voltage across a diode to the resulting change in plate current is known as A.C. plate resistance.



$$r_p = \frac{\Delta V}{\Delta I}$$

Note : The A.C. plate resistance at operating point P can be found by considering small equal changes of plate voltage on either side of the operating point (*i.e.*, AB = A.C.)

Change in plate voltage = BC

Change in plate current = YZ

$$\therefore \text{ A.C. plate resistance at } p,\ r_p = \frac{BC}{YZ}$$

2. At $V_p = V_1$, almost all the electrons constituting space charge are collected by plate and space charge is eliminated. At this point rate of collection of electrons reaches its maximum value and it equals the rate of emission of electrons by cathode. Therefore, if plate voltage is further increased plate current becomes constant.
3. The only way to increase the plate current is to increase the rate of emission, *i.e.*, to increase the temperature of cathode. Since the plate current, in this region, may be limited by temperature of cathode, this is called **temperature limited region**. In this region plate current is independent of plate voltage and varies with temperature T of cathode as

$$I = cT^2 e^{-b/T} \qquad \ldots(ii)$$

where *c* and *b* are constant for cathode.

Equation (ii) is simplified form of Richardson equation for this case.

Plate resistance of diode—The plate current (I_p) varies as the plate voltage (V_p) is changed, therefore, a diode offers internal resistance which is known as its plate resistance.

Note : Mainly negative space charge is responsible for the plate resistance of diode. As it is a non-ohmic resistance its value is not constant and it differs at different operating points.

Types of Plate Resistance

(i) **D.C. plate resistance**—The ratio of total D.C. plate voltage across diode to the resulting current is known

We are going to compare the sensitivity at 1000 kHz, with a high quality receiver, with L = 200 µH and unloaded Q = 1000.
This receiver has optimal matched diode and load resistor.

The values of the simple receiver can directly be entered in calculator 2.
We don't have to calculate first the optimal load and diode, because this receiver isn't using optimal load and diode.
So, in calculator 2 we fill in:
f = 1000 kHz
L = 200 µH
Q = 200
Is = 400 nA (at 25 ºC).
RL = 47 kΩ.

Fill in the values of the other receiver, in calculator 1:
f = 1000 kHz
L = 200 µH
Q = 1000
Click on "calculate calculator 1" (The optimal load resistor and diode are now calculated).
Click on "calculate calculator 2" the differences between both receiver are calculated.

Result: the simple receiver has 29 dB less output power at weak signals.
The simple receiver has a loaded Q of only 60 (at weak signals).

At strong signals, the loaded Q even reduces to 17, but with such a low sensitivity it's unlikely to get a strong signal across the LC circuit.

We note from fig. (i) that curves coincide at low voltage (in space charge limited region) but saturation current increases as shown by portion CD of characteristics.

(2) **Low voltage characteristics**—The plate current is practically zero at zero plate voltage (however, plate current is not exactly zero). If we measure plate current using a micro-ampere plate current of the order of few micro-ampere occurs. This is due to "the fact that even at zero plate voltage a few electrons may have sufficient K.E. to reach the plate and constitute a plate current of few micro-ampere." From the fig. (ii) the plate current becomes exactly zero at a particular negative value V_C of plate voltage, called **cut-off voltage.** The max. kinetic energy of emitted electron is related to V_C as :

$$K_{max} = e\,V_C$$

Important Points

1. If we increase plate voltage from zero volt to a value V_1 in steps, more and more electrons are attracted from space charge to plate (though at the same time same no. of electrons are emitted from cathode so that space charge is maintained) so plate current increases. As the plate current is limited by space charge, this region of curve is called **space charge limited region**. In this region, plate current is related to plate voltage by equation.

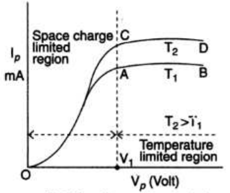
$$I_p = K \cdot V_p^{3/2} \quad \text{...(i)}$$

where K is a constant.

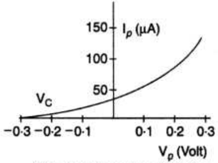
Equation (i) is known as **Child-Langmuir's law.**

V-I characteristics of Diode

(1) **High voltage characteristics**—Fig. (i) shows high voltage plate characteristics when plate current is measured in (mA). Keeping temperature of cathode constant, say (T_1) (*i.e.* filament current constant) when plate voltage increased from zero volt in steps and corresponding value of plate current i_p measured in m.A. the variation occurs as shown in curve OAB. If the cathode temperature is increased from T_1 to T_2.



(i) High voltage characteristics



(ii) Low voltage characteristics

Fig. : V-I characteristics of diode.

Crystal Set Analysis
by Berthold Bosch, DK6YY
http://www.oldradioworld.de/gollum/analysis.htm

(First published in German in 1993/94; see references at end. Updated 12/03/2002)
(See some of Berthold Bosch´s realized xtal sets here)

Contents: 1.Voltages and Powers in Set, 2. Antenna/Earth as Signal Source, 3. Set with Parallel- Tuned Circuit, 4. Diode Properties, 5. RF Matching, 6. AF Matching, 7. Computer Simu- lations, 8. Series-Tuned Circuits .

Since my schoolboy days I have been fascinated by crystal radio reception: radio in its most basic form. However, in many cases I was not really satisfied with what I read on the subject in the literature. In the treatises I came across, the descriptions often remained rather vague, presenting little convincing foundations. Partly they were rather speculative and even presented contradicting conclusions. For this reason I found it advisable to carry out my own investigations. My intention was to obtain more quantitative results, for example as regards the best diode and the understanding of the obvious interdependence between the radio-frequency (RF), audio-frequency (AF), and DC subcircuits, what it meant for an optimum design. In the following I present results obtained over the years. Only medium-wave reception is considered.

1.Typical Values of Voltages and Powers in Set

Let us first see of what order of magnitude the RF and AF voltages and powers are which we have to deal with.

According to amplitude-modulation theory, the AF power contained in the total AM signal of power PRF is given by m2

PRF / (2+m2) where m is the modulation factor. If we assume m=0.5 we thus have 11 percent of AF power in the AM signal. Sometimes broadcasting stations use modulation factors of up to m=1 (100 percent) which then causes a correspondingly higher AF power component. At my urban location, in the West of Germany (Ruhr District), the strongest station (15 km away) produces an electric field strength of 0.18 V/m and, with my antenna and earth arrangement, an RF power of about 3 mW is available in the crystal set. Hence 330 µW of AF are contained in the RF if we assume m=0.5. A practical (linear) diode detector coupled to a tuned circuit delivers 70 to 80 percent of this to the AF load. This means that ideally I can expect about 240 µW of AF being available from my local station, sufficient for moderate operation of a loudspeaker.

In the crystal set that I am going to investigate (Fig. 2 below) I measured the following RF voltages across the tuned circuit when RF and AF matching existed (Secs. 2, 5 & 6):
a) Tuned to the local station (WDR 2, 720 kHz, 200 kW, 15 km away):8.9 V

b) From of my "district station" (DLF, 549 kHz, 100 kW, 35 km, field strength 40 mV/m, 0.2 mW of RF power):2.3 V.

c) At night - with a wave trap for the local station - more than a dozen stations appear from all over Europe with 1 to 5 mV/m, producing 130 mV across the circuit as a mean value (0.5 µW of RF). Such low voltages will move the working point over only a rather limited part of the diode characteristic where the relative curvature is low. Consequently, the detector efficiency now drops to below one percent.

d) Good headphones produce an audible signal down to 10 pW of applied AF power.

plate current. These electrons flow through the external circuit and finally return to the cathode, thus making up the supply of electrons lost by emission. On increasing plate potential more electron will gain sufficient kinetic energy so as to reach anode, hence the plate current increases.

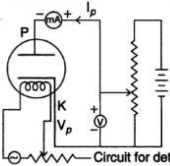
Important Points to Note

- The current flows in the diode only when plate is made positive relative to cathode.
- No current can flow when plate is negative relative to cathode.
- Within a diode electrons can flow only from cathode to plate.
- Due to unidirectional conduction property the diode acts like a valve.
- The diode automatically starts conduction when the plate is positive or stops conduction when the plate is negative (due this property) the diode may act as a rectifier, converting A.C. into D.C.)

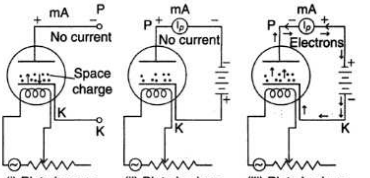
Characteristics of Diode

Characteristics of vacuum diode can be best studied by finding the relation between plate voltage and plate current for a given cathode temperature.

The circuit for determining the plate characteristic of an indirectly heated vacuum diode is shown in fig.



Circuit for determining plate characteristics of diode



(i) Plate is open (ii) Plate is given –ve potential (iii) Plate is given +ve potential

(1) **Anode (plate) at zero potential relative to cathode**—This situation is shown in fig. (i)—The emitted electrons do not have sufficient kinetic energy so as to reach the anode. However, a few electrons may reach the anode on account of their kinetic energy, constituting negligible current. The emitted electrons accumulate near the cathode and form a cloud of electrons. This is known as space charge. At a certain stage the number of electrons forming the space charge becomes constant for a given operating temperature. This space charge becomes a source of electrons that can be attracted to the plate, if it is at a positive potential.

(2) **Anode (plate) at negative potential relative to cathode**—This situation is shown in fig. (ii). The emitted electrons are repelled back due to retarding potential of anode. For a particular negative potential of anode plate current may be zero.

(3) **Anode (plate) at positive potential relative to cathode**—This situation is shown in fig. (iii). The electrons constituting the space charge are attracted to the plate. This flow of electrons from cathode to plate is known as

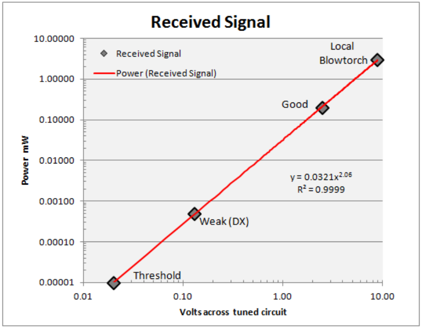
Employing a signal generator and using a sensitive diode (see below) I found that an RF power of about 10 nW is required to generate this 10 pW of lower-limit AF. The detector efficiency has at this very low RF level thus fallen to a mere one per mille. Obtaining an RF power of 10 nW in my set requires a field strength of about 0.3 mV/m. According to estimates based on groundwave propagation theory, a 1000 kW transmitter operating near 1500 kHz should generated 0.3 mV/m at a distance of 190 to 200 km; the electric field strength is roughly proportional to Radicle (PTX)/(f2d2), where PTX = transmitter power, f = frequency, and d = distance. The particular example is chosen because at 1440 kHz I can during the day just hear the signal of RTL Luxembourg, being 195 km away and reported to radiate 1200 kW. The voltage measured across the tuned circuit was 40 mV in this case. To be able to receive RTL I carefully have to suppress the local as well as the district station.

e) When I connect an AF amplifier to the crystal set, a number of stations located about 150 to 250 km away can additionally be heard via groundwave propagation in the daytime. The diode thus provides (some) detector action at RF levels even lower than 10 nW. But the AF power generated is then too small to produce an audible signal in the phones directly.

The above numbers show the considerable variations in the RF voltage generated across the tuned circuit. Hence it is not surprising that a crystal diode found to be best suited for DX reception is not necessarily the optimum choice for achieving best loudspeaker operation from the local station. But before I present results on diode behaviour I am going to describe the circuit I employed. Let us start with antenna and earth as an integral part of the total circuit.

Ed Note: Evaluation of Received power from above:

		threshold	weak	medium	strong
RF power	W	1.00E-08	5.00E-07	2.00E-04	3.00E-03
RF Field	mV/M	0.3	2.5	40	180
RF voltage	V	0.02	0.13	2.5	8.9
	mV	*20.0**	130	2500	8900
		(* 40 in text: 20 provides superior fit)			
	Antenna:				
	Inverted L design, 43m (140')				
	Height = 10m (32')				
	Earth Rg = 210 Ω	Counterpoise Rg = 25 Ω			



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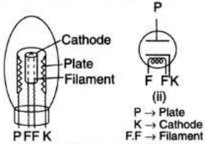
Vacuum Diode or Diode Valve

A vacuum diode consists of two electrodes : a cathode; and an anode enclosed in an evacuated glass tube. Its operation is based upon thermionic emission, it is also called thermionic diode.

Anode (or plate)—The anode is a hollow cylinder made of Ni or molybdenum and surrounds the cathode.

Cathode—A cathode may be directly heated or indirectly heated type. Usually an indirectly heated cathode is provided (except in high power application).

As shown in fig. (i) an indirectly heated cathode is in the form of nickel cylinder coated with oxides of barium or strontium. Inside the cathode, a heater filament of tungsten is inserted.



(i) Construction of diode
(ii) Symbol of diode

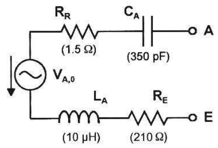
Fig. (ii) shows it symbol :

Operation of diode—When electric current is passed through the diode it emits a large number of electrons. These emitted electrons may be accelerated or retarded by applying positive or negative potential to anode relative to cathode.

when the tube is in operation. It can just be seen glowing orange in the diode portion of the tube. On one side is the diode with the triode on top; on the other is the pentode, all of which are clearly displayed. The tube cost $3.10, an excellent price considering all the experiments that can be done with it.

The center tap of the input transformer is not used in this circuit. The triode has grid leak bias. It is not possible to measure this bias with a DMM because of the large grid resistor, but the plate current is only about 280 µA. The diode and triode are referred to the negative end of the filament, but cathode bias is provided for the power pentode. The bias in operation is about -9.3 V, giving a plate current of about 5 mA and a screen current of about 1 mA. The output transformer can be any transformer on hand. I used the P-T31 that has been used in other places, which matches 8Ω to 5 kΩ. The load resistance should preferably be about 12 kΩ. However, it drove a good speaker at considerable volume with a rather small input. A three-tube battery radio could be made with this tube, an IF amplifier stage, and a pentagrid converter, a tube type which will be treated next.

2. Antenna and Earth as Signal Source



(Fig. 1: Equivalent circuit of antenna/earth combination.)

I use an inverted L-type antenna of 43 m length, about 10 m above ground. The earth connection is provided by three metal rods of 2 m length each, driven into marly, i.e. a not particularly well conducting, soil. The antenna/earth combination can be represented by the equivalent circuit shown in Fig. 1, giving measured values for the various elements. The antenna capacitance is denoted by CA, the inductance by LA, RE is the earth loss resistance, RR the radiation resistance, and VA,O the antenna source voltage. The two last elements increase in value with the antenna's height and length. For the source voltage I measured a value of 1.6 V, using a selective RF voltmeter. The knowledge of this quantity, which is produced by the strong local station, permits to easily determine the earth resistance. For it 210 ohms were obtained, a relatively high value. Reducing it by installing a better ground system would pay high dividend. Note added in 2002: Meanwhile I installed an extensive counterpoise net in the garden as earth terminal. This reduced the earth resistance to about 25 ohms, with an associated marked increase in available RF power.

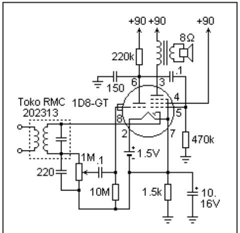
Maximum power is transferred to the load, i.e. from the antenna to the crystal set connected to A-E, when we arrange for impedance matching and for resonance in the resultant antenna/earth series circuit. The set will presents, in general, an inductance which is too small for achieving resonance in the antenna circuit. Therefore, an additional coil must be inserted (Fig. 2). The resistance of about 400 ohms in the series-tuned circuit (200 ohms source resistance plus 200 ohms resistance of set when matched) yields a Q-factor of only 4. But this is still helpful regarding sensitivity, but also selectivity (sharpness of tuning), since the delivered current (voltage) is increased by this factor of 4, meaning 16 times in power.

3. Crystal Set with Parallel-Tuned Circuit

As a sort of "standard set" I used and investigated the popular arrangement shown in Fig. 2 which employs two tuned circuits. The inductance LC couples the antenna to the coil of the tuned circuit, the degree of the variable coupling chosen so that matching is achieved. The fixed L1 has a somewhat larger value than required for tuning the antenna/earth circuit of Fig. 1. The variable capacitor C1 is then used for tuning to resonance. The numbers for L1 and C1 apply to my particular case. C3 serves as AF storage capacitor for obtaining a maximum of AF amplitude at the phones, and it additionally provides a short for the RF. In practice, however, it often can be omitted without audible drop in AF. - To be able to properly match the diode detector to the parallel-tuned circuit, the detector branch is hooked up either to the top of coil L2 (wound with Litz wire onto a suitable ferrite rod) or to one of 11 taps provided on it. In this way the diode can be connected, via a switch, to 12 resistance values along the tuned circuit. Such a fine adjustment was required for the investigations reported in Sec. 5. The unloaded tuned circuit has a resonance resistance of 105 k ohms (at 1000 kHz), which drops to 52 k ohms when matched to the antenna/earth. (These are not particularly high resistance values because of the many leads from the taps to the switch.) According to the switch position chosen the diode can so be connected to 12 resistance values that vary between 52 k ohms and 100 ohms. When the diode is set to the tap that provides matching the total resonance

circuit at the left. AVC is essential when diode bias is used, to keep the bias voltage in the proper range. The audio signal, of course, rides on the bias connection. Diode bias is not suitable for high-mu triodes such as the 6AV6 or 6AT6, since they are very sensitive to the bias level. The 6R7 is a medium-mu triode (μ = 16). The 6SR7 is a single-ended equivalent, and will also work in this circuit. The signal generator should be adjusted to peak the output, at 455 kHz. I obtained an audio output of 13 V peak-to-peak with a 30% modulated AM signal of 0.2 V peak-to-peak, for a gain of about 289. The tuned circuit provides a large part of this gain.

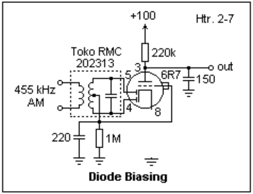
The 6AQ6, 6AT6 and 6BF6 are all miniature dual-diode triodes with the same basing as the 6AV6, the 6BF6 being medium-mu and more appropriate for diode bias and transformer coupling. The 6SQ7, 6SR7 and 6ST7 are similar octal tubes, again all with the same basing. This was a very popular and useful tube type, with many options available.



Detector - Audo Amplifier

The 1D8-GT diode-triode-pentode was designed to provide a complete audio system for a battery-powered radio. The circuit is shown at the right. The filament takes 0.1 A at 1.4 V, easily supplied by a D cell. The B+ supply is 90 V, for which batteries were available, consisting of 60 small cells. This circuit has a drain of less than 7 mA, which is quite acceptable, and hardly more than a transistor radio with the same audio output. The filament is hard to see the plate of the 6AT6. The RF ripple was negligible, but there was some 60Hz pickup, not surprising with a high-gain breadboarded circuit out in the open. It was interesting to show the AM signal and the audio output simultaneously on the oscilloscope.

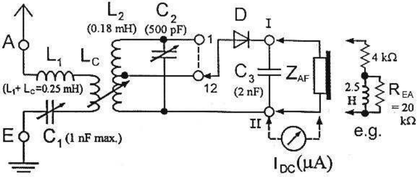
The circuit will function without the tuned circuit, as you can easily demonstrate. The signal generator output must then be increased, because the resonance gain is considerable. The circuit will also function with both diode plates connected to one side of the secondary, and the 1M potentiometer and filter capacitor to the other side. In this case, the ripple frequency will be halved, but the output voltage will be doubled. There is not a great advantage in using a full-wave detector. Look at the DC voltage across the potentiometer as the input amplitude is changed. This voltage, negative with respect to ground, can be used for AVC, automatic volume control, as is discussed elsewhere. With the full-wave detector, it varied from -0.5V for 10V p-p audio output to -1.0V for 30V audio output. In practice, the AVC voltage is further filtered to remove the audio.



Diode Biasing

The rectified signal voltage is of the correct polarity to provide grid bias for the amplifier tube in this circuit. Therefore, the grid of the triode can be connected directly to some point on the diode load resistor that provides the proper bias, in place of the volume control potentiometer and coupling capacitor shown above. This arrangement is called *diode bias*, illustrated by the

resistance then drops to 26 k ohms. - On the right in Fig. 2 the equivalent circuit of the headphones is given which we require later on.

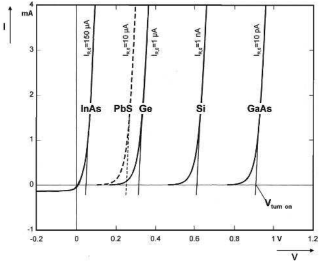


(Fig 2: Wiring diagram of crystal set with parallel-tuned circuit and tuned antenna/earth.)

We have now the task to match (1) on the RF side the diode branch, via the tuned circuit and the antenna coupling, to the antenna source, and (2), regarding the AF, the phones or the loudspeaker to the diode. Both procedures are strongly interrelated. But first we must find out more about the diode and its dynamic resistance.

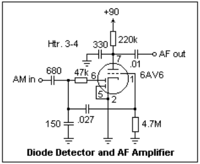
4. Crystal-Diode Properties

All semiconductor diodes principally show a dependence of current I on applied voltage V (static characteristic) like $I=I_{R,0}*\exp(V/nV_T) - I_{R,0}$, where $I_{R,O}$ denotes the reverse saturation current at high negative voltages, V_T the temperature voltage of 26 mV at room temperature, and n an ideality factor between 1 and 2.

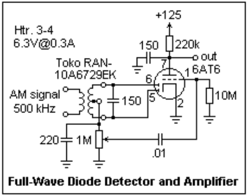


(Fig. 3: Theoretical current-voltage characteristics of various p-n diodes; equal diode areas assumed.)

Reverse breakdown is neglected. The relationship applies to both to p-n diodes and to Schottky diodes. However, the composition and physical meaning of IR,O differs completely for these two diode types (diffusion current resp. field current). Important is the fact that the exponential rise of the diode current occurs the faster, i.e. the turn-on voltage (as a kind of threshold) becomes the lower, the higher the saturation current IR,O is. Calculated characteristics for p-n diodes made from germanium, silicon, two modern compound semiconductors and from galena (PbS) are given in Fig. 3, assuming equal diode areas. The associated reverse currents are stated. The difference in diode behaviour is caused by differing electronic material properties (band gap). Real curves, in particular of the natural crystals, are flatter because of parasitic elements, mainly of diode series resistance. Regarding low turn-on voltage a galena p-n crystal diode is theoretically even slightly better than one made from germanium.

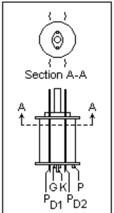


A detector-amplifier circuit is shown at the left. This circuit is fed by a signal generator at AM in, and the audio output is taken from AF out. The amplifier uses grid-leak bias, which effectively clamps the peak of the audio wave at 0 V. The stronger the input signal, the more negative the grid bias, which makes the most of what is there. With an AM input of peak value 0.95 V and modulation amplitude 0.37 V, the peak to peak output was 3 V, for a gain of -8, which is quite good. I used a radio frequency of 2 MHz, and modulation of 1 kHz. At lower radio frequencies, the filtering of the output becomes progressively worse; this could be optimized if desired. This detector gives good fidelity with high sensitivity.



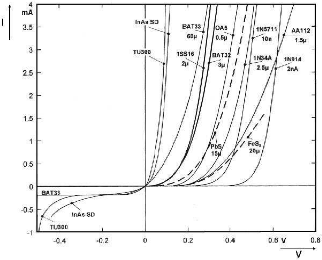
A full-wave diode detector is shown at the right, using transformer input. The 6AT6 is similar to the 6AV6, but its mu is 70 instead of 100. The transformer is a Toko RAN-10A6729EK, available from Digi-Key. It is specified as 0.63 mH, and for 200 kHz. Adjust the signal carrier frequency until the tuned circuit resonates; in my case, this was 500 kHz. For an input signal from the signal generator of 0.18V peak-to-peak (modulation amplitude 0.04V), I got 26V peak-to-peak at

The typical superheterodyne receiver used a diode detector, followed by an audio voltage amplifier. These functions were usually combined in a single envelope, the diodes and the amplifying triode or pentode sharing the same cathode. A very good tube of this type was the 6AV6, whose structure is shown at the right. The diode plates are at the top of the envelope, with the cathode easily seen between them. The IF output was usually from an air-core, slug-tuned transformer, making it easy to use full-wave rectification. Only one of the diodes could be used for half-wave rectification, which was usually quite adequate. The triode is a high-mu triode with a maximum plate voltage of 300 V, and maximum plate dissipation of 0.5 W. The maximum cathode current is not specified, but should not exceed a few milliamperes. The tube has an unusually high plate resistance, and the amplification factor is over 100. It is a good triode for a voltage amplifier. My example gave g_m = 1.3 mS, r_p = 67k, μ = 90 at about 500 μA plate current, and V_g = -1.0 V. The transconductance varies strongly with plate current, reaching 2.2 mS at 1 mA, where μ = 116.



6AV6 Structure

For Schottky diodes principal curves like those in Fig. 3, relating to a particular crystal material, cannot be given. In their case the characteristics strongly depend on the kind of the metal electrode and on the processing parameters. But measured curves for Schottky as well as for p-n diodes are given in Fig. 4, including fairly good examples of galena (PbS) and iron pyrite (FeS) detectors; see also the Table below for identifying the diode types. Again one notices the relatively low turn-on voltage of the two natural crystals which equal or even are below that of the germanium diodes 1N34 and OA5, thus being well suited for low-level detection. (Fig. 4: Measured current-voltage characteristics of various semiconductor diodes (reverse currents indicated))



Very interesting is the performance of modern low-barrier Schottky diodes made from silicon, like the NEC 1SS16 (almost identical: 1SS99, BAT32, BAT63), which show turn-ons at 0.15 to 0.18 V. And, indeed, they show superb performance at low levels. One should expect that the InAs Schottky diode (which was specially made for my experiments) and the TU 300, a backward diode made by

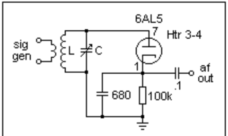
Siemens, would according to the curves shown be even more sensitive detectors. But this is not the case.

As mentioned, a low turn-on voltage is inevitably associated with a high reverse current. This current reaches values of a few hundred µA for the InAs diode, as also for the TU300 and the Schottky diode BAT33. If the reverse current, i.e. an unwanted back current, reaches such high values we have strong counteracting effects, and ultimately the detector action disappears completely. Anticipating the results of computer simulations described in Sec. 7 one can state that diodes like the 1SS16 show the optimum relation between low turn-on and still acceptable reverse current, thus making them the best choice of presently available diodes as regards detector sensitivity.

To show and compare the capability of various diodes the Table summarizes values of measured AF voltages and of rectified currents, for 1 and 100 µW of available RF power. A power of 1 µW is in my set typical for DX stations at night, and 100 µW for stations 30 to 50 km away. As is seen the 1SS16 leads the field. - For 3 mW of RF (my local station) I obtained with a 1SS16 a DC current of 715 µA, which increased to 1.85 mA in the short-circuit case (AF/DC load = 0), and to 2.95 mA when under these conditions the set was retuned.

positive in a resistance-coupled amplifier, disturbing the bias. This is not a maximum in the same sense as the maximum plate dissipation or maximum cathode current are, since damage to the tube is not in question. There is no trouble in the grid leak detector, and values of several megohms are permissible.

There is no analogy to the grid-leak detector with FET's, which otherwise behave pretty much like vacuum tubes, because the leakage current is far too small for the desired behavior. As a gate-bias detector, however, the FET is suitable, and a BJT biased near cutoff makes an excellent detector, as we saw in the page on Amplitude Modulation.



Diode Detector

A diode detector, using a standalone diode, is shown at the left. The output is the same amplitude as the modulation on the carrier that is input, very closely, so the detector could not do any better. Often, the diode was included in the first AF amplifier tube, sharing the same cathode, and dual diodes were also available for full-wave detection, which doubles the output. If you have a tuned circuit at this point, it would be easy to test a diode detector, and high voltages are not required, just the heater power. A solid-state diode would work the same way, but the 0.7 V drop would make it less sensitive, and it would not even respond to weaker signals. Using a Ge diode (1N34A) would help, but not eliminate, the problem. The thermionic diode is ideal in this application.

coupled by about 9 turns slipped over the loopstick, providing an AM signal about 30% modulated with 1 kHz, in the broadcast band. The output was roughly constant, no matter what the plate voltage (45 to 105 V), input amplitude, value of grid resistor (from 1M to 3M) or grid capacitor (100 pF to 220 pF). The grid bias with no signal was about -0.5 V with a 1M grid resistor.

The circuit works by using the small grid current when the grid is slightly negative. This grid current increases rapidly and very nonlinearly, so square-law detection is possible. The grid resistor establishes a quiescent point around which the signal oscillates. The rf part is bypassed through the grid capacitor, while the af part remains and causes a voltage drop across the grid resistor, which is then amplified in the usual way. With a 100k plate resistor, the output af voltage could approach 1 V peak-to-peak. Another stage of audio amplification would give a quite satisfactory result, but I was hoping for one tube. (The 12AX7 has two triodes in the same envelope--if such a tube were used, the receiver would be, strictly speaking, single tube.) With most other kinds of detectors, such a hope would be quite vain, it must be admitted.

If the input signal amplitude is increased, the grid bias starts to decrease because the grid becomes positive on the peaks, and the detection becomes a little more linear. If the grid is biased to cutoff with a C supply, then there is clear linear detection, since the signal is effectively rectified. Such a circuit is called a grid-bias detector, and the signal amplitude must be restricted so that the grid does not go positive, when distortion would result.

You may notice in the tube manuals that a "maximum grid resistance" is given for a tube; for the 6J5, it is 1M. The reason for this is positive ion current, which may drive the grid more

Type	Kind	V_{AF} / mV	I_{DC} / µA	V_{AF} / mV	I_{DC} / µA
1SS16	Si SD	36	10.5	360	152
1N34A	Ge pn	26	6.0	312	121
PbS Det.	Galena	25	6.5	301	115
AA112	Ge pn	24	5.5	305	118
OA5	Au/Ge pn	22	4.5	285	120
1N5711	Si SD	16	2.5	260	80
FeS_2 Det.	Iron Pyrite	12	2.0	235	85
1N914	Si pn	2.5	0.2	320	110
BAT33	Si SD	1	0.5	35	12.5
In As	Experim. SD	0.5	0.2	29	10
TU300	Si BW D.	<0.1	<0.05	65	21
		$P_{RF,0}$ = 1 µW		$P_{RF,0}$ = 100 µW	

(Table: Measured values of AF voltage (across phones of 4 k at DC) and of rectified DC current for various diodes and two levels of RF power)

Sometimes a DC bias from a battery is applied for shifting the operating point of the diode closer to the turn-on voltage and so improving the detection efficiency. By this method the AF voltage obtained can be increased, for example when using the 1N5711 and the 1N914 at low RF levels. The 1SS16 group of diodes, however, hardly gains from a DC bias. Only at RF powers below about 200 nW I was able to measure a certain rise in AF voltage. At the lowest detectable RF level of 50 nW (Sec. 1), the AF voltage increased by 20 percent (i.e. power by 45 percent) when the optimum bias was applied. But this effect was measurable only, being still too small to be noticed by the ear.

Note added in Jan. 2002: Backward diodes (BWD), like the TU300, are good detectors at extremely low RF signal levels, below about 1 nW with associated voltages of only a few mV. This is due to the relatively sharp bend in the BWD characteristic at zero volts. The generated AF signal is, however, too small for operating phones directly and calls for an AF amplifier. Then stations can be copied which are not heard when in such a set-up with AF amplifier a "normal" sensitive diode, like the 1SS16, is used instead of the BWD.

5. RF Diode Resistances and RF Matching

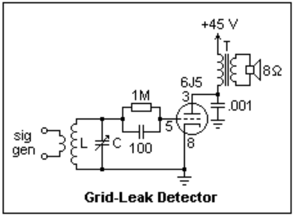
For achieving best performance it is required to RF match the diode to the tuned circuit. The dynamic resistance of the diode depends on the amplitude of the RF voltage applied to it, and on the kind of AF load impedance.

In AM tube radios the detector diode operates at a high level (linear detection) and has a load consisting of a large (ohmic) resistor shunted by a small capacitor. Calculations show that in this case the RF diode resistance, as presented to the tuned circuit, is roughly half of the ohmic load resistance. In a crystal set the calculation is somewhat more complicated since there the RF voltage on the diode is generally lower and the diode load is more complex (see equivalent circuit of phones in Fig. 2). Hence I preferred to measure the RF diode resistance RD. The measurements were carried out under actual working conditions using a signal generator. Figs. 5(a) and (b) show the results obtained for high-impedance phones with 4 k ohms DC resistance and for low-impedance ones with 120 ohms, respectively. The RF frequency used in these measurements was 1000 kHz, the modulation frequency 1 kHz with a modulation factor of 0.4 (given by the signal generator). Figs. 5 give the measured diode resistances, as a function of the RF power applied, for an 1SS16 (also some in parallel), a silicon p-n diode 1N914, and for natural galena as well as carborundum (silicon carbide; SiC) crystals. The diode circuit was in turn connected to the various taps on the coil L2 .When the RF voltage measured across the tuned circuit dropped by a factor of radicle (2)=1.41 compared to its value without diode, matching was achieved. Then the RF diode resistance equalled the RF resistance of the tuned circuit at the tap point. To avoid

Unfortunately, I do not have any actual circuits that used a 1V6, but the amplifier shown at the left may be studied. It gave a gain of -14 with a plate current of 357 μA, so the effective transconductance was 298 μS. I substituted a 100k plate resistor, but the gain only increased marginally, to about -17. By experimenting, you may be able to get more gain from the tube. At low plate currents, the amplifier oscillated in an odd manner, and there were other peculiarities seen while fiddling around, but I did not take the time to track them down.

The Grid-Leak and Diode Detectors

Detectors are discussed in Amplitude Modulation, where the two types of square-law and linear are explained, and the grid-leak detector is mentioned. A circuit using a 6J5 triode is shown at the right. The audio output transformer T can be replaced with a 10k or 100k resistor, and the output observed with an oscilloscope. I could find only a 1k to 8Ω transformer, really designed for use with transistors. The circuit works, but I was not impressed by its output. Asking for loudspeaker output from a detector is being very optimistic, but I had some hopes. Perhaps a more suitable transformer would give better results.



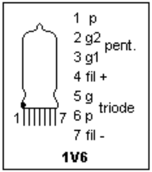
Grid-Leak Detector

The tuned circuit can be one from a crystal set, with a loopstick and variable capacitor. The signal generator is

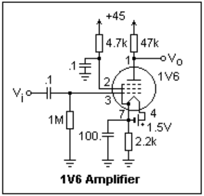
These tubes make vacuum-tube experiments very easy without requiring high-voltage supplies, only what is found in a normal transistor lab.

Sub-Miniature Tubes

The smallest regular electron tubes were the *subminiature* tubes, designed for battery-powered portable apparatus, from hearing aids to radiosonde transmitters. Typically, the lead wires were brought out through the press seal without a base, but there was also a subminiature base used for a few tubes. As an example, you can work with the 1V6, a pentode-triode with a filament taking 40 mA at 1.25V (the filament supply can be a D cell). The tube is 40mm x 10mm x 7mm, approximately. The maximum plate voltage is 45V, and typical plate currents are less than 1 mA.



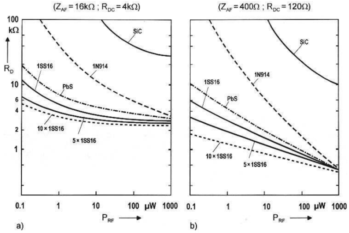
A sketch of the 1V6 and its connections is shown at the right. On the press seal, there is a red dot that identifies what I have called pin 1. The filament is located at the center of the tube, with the pentode on one side and the triode on the other. The triode shows a μ of about 10, r_p = 18.6 kΩ, and g_m = 440 μS. The pentode has g_m as high as 694 μS above 500 μA plate current, but it drops considerably at lower plate currents. The two devices can be tested as usual, taking care not to exceed 45V or 1 mA. Note that the pentode and the triode share the same filamentary cathode.



an error one must readjust the coupling to the antenna when the diode is connected to the first found (V/1.4) tap point and then repeat the search for the now somewhat altered (V/1.4) tap. A second iteration further improves the result, but not much.

As in principle to be expected from the characteristics, the diode resistances vary rather widely, from some 100 ohms to some 10 k ohms, with lower values obtained when the DC resistance of the phones is low. The galena detector shows values only moderately higher than those of a single 1SS16. The silicon diode 1N914 presents high values due to its high turn-on, which even more applies to carborundum.

(Fig. 5:Measured RF diode resistances versus available RF power: (a) for high-impedance phones (4 k ohms at DC), (b) for low-impedance phones (120 ohms at DC)

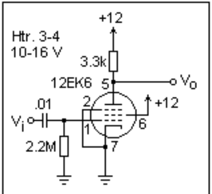


The data obtained then indicate that the optimum tap position on coil L2 (for matching) depends on the diode type, the strength of the received station, and on the DC resistance of the phones. The larger the value of the diode resistance is, the higher must the tap position be up the coil. Sometimes it was suggested in the literature to have a fixed tap at a point of about 1/4 to 1/3 of the windings counting from the earth point. In the present case the tuned circuit has a resistance of approximately 6 k at the 1/3 tap point. As Fig. 5a shows, this indeed is a rather good choice for a galena detector when high-impedance phones are used and weak stations received. Impedance matching requires that the reactances of source and load cancel out. But in our case the resistance of the tuned circuit has no reactive part at resonance, and the reactance of the diode, caused mainly by the diode junction capacitance of at most a few pF, can be neglected.

Connected to a particular tap, the diode resistance is (auto-)transformed up and appears in parallel to the resonance resistance of the tuned circuit. This means that not all of the available RF power reaches the diode since a reasonable fraction of it is dissipated in the resistance of the tuned circuit. In order to really transfer the maximum of power from the antenna to the diode branch, the diode (of generally low resistance compared to that of the tuned circuit) should be connected untapped to the top of the coil L2. This, however, strongly reduces the selectivity of the set and requires a readjustment of the coupling of the tuned circuit to the antenna. With high incident high RF power (and/or low impedance of the phones) the tuned circuit can, under these conditions, become loaded to such an extend that variations of the capacitor C2 have no tuning effect any longer, which means that C2 is obsolete and can be omitted. The diode circuit is then aperiodically coupled to the (tuned) antenna circuit, while the coil L2 merely acts as the secondary winding of the transformer which matches the diode to the antenna.

6. AF Matching

The 12EK6 (also known as 12DZ6 and 12EA6) is a screen-grid pentode for low-voltage use. A voltage amplifier using this tube is shown at the left. This tube was generally used as an IF or RF amplifier, so the load would have been a resonant circuit, not a resistor, giving much higher gain. However, the resistive load is easy to experiment with. There is no voltage to waste for cathode bias, so grid-leak bias is used. When a signal is applied, the bias will adjust itself so that the top of the input waveform is clamped to 0 V. Vary the input amplitude to verify this. This circuit works with an input of up to about 2 V peak-to-peak, perhaps best for an input of around 1.0 V p-p. The output voltage is then about 5.8 V p-p, the plate potential is 6.0 V, plate current 1.8 mA, and gain about -5.5, corresponding to a transconductance of 1.8 mS, about right from measurements of the characteristics and the position of the operating point in this case. The transfer characteristic is rather curved, leading to considerable distortion for larger amplitudes. Nevertheless, it is remarkable that the circuit works so well at such a low voltage.



Low-Voltage Amplifier

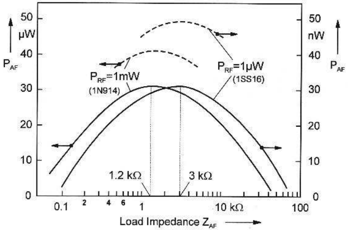
As a triode (plate, suppressor and screen connected together), the 12EK6 has $\mu = 8.5$, $g_m = 6.3$ mS and $r_p = 1.35$ kΩ. The transconductance varies considerably with plate current, which gives rise to distortion for large signal amplitudes.

current, which was stepped up by a transformer and then rectified to DC again (sometimes by contacts on the vibrator) for the B supply. Vibrators were electrically very noisy, and had short lives, but were widely used until solid-state equivalents became available. They came in cylindrical metal cans, and looked like electrolytic capacitors. The 0Z4, mentioned elsewhere, was a rectifer specifically for vibrator supplies. Another possibility was the *dynamotor*, a motor-generator with a single rotating armature and brushes, which could supply more power than a vibrator and was more reliable. However, they were too expensive for general commercial use, though widely used in the military.

When 12 V became the automotive standard, vibrator supplies or their solid-state equivalents, could still be used. Tubes with 12.6 V heaters were already common. Many had the heater center-tapped, so they could be used equally well on 6.3 V. However, 12 V is high enough to be used directly as the plate supply if the tubes are specially designed. A range of tubes was produced that could be used on 12 V only, without any high voltage at all. We have already discussed the 12DL8 and 12K5 space-charge-grid tetrodes, and how the space-charge grid allows larger plate currents for small plate voltages. There were approximately 15 tube types designed for 12V use, of which only three, the 12DL8, 12DS7 and 12K5 are space-charge grid tetrodes, behaving like triodes. The other tubes are of conventional construction, in general having low plate currents, less than 1 mA in many cases. These include twin-diode triodes (12AE6, 12FK6, 12AJ6), sharp cutoff pentodes (12AF6, 12BL6, 12CX6, 12EK6), remote cutoff pentodes (12CN5, 12DZ6), twin-diode remote cutoff pentode (12F8), pentagrid converter (12AD6), pentagrid (variable gain) amplifier 12EG6, and a twin-diode power screen-grid tetrode, 12J8. The last tube can deliver 20 mW output power, half of what the 12DL8 can do.

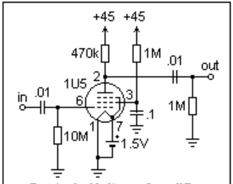
If a crystal ear phone is used or the diode detector is followed by an amplifier (generally of high input impedance) one has to design for maximum voltage at the detector output. Here, we rather have to deliver a maximum of power to the phones. Hence the impedance of the phones (or the speaker) as the AF load should have such a value that a maximum of AF power is transferred to it. The AF source resistance RG is at low RF levels (square-law detection) approximately given by the reciprocal of the slope of the diode characteristic at the operating point. At higher RF levels (linear peak detection) it is determined by the current spikes flowing through the diode. In so far, RG nearly equals the diode resistances as shown in Figs. 5. The tuned circuit presents an AF short.

I determined the equivalent circuit of a pair of high-impedance Telefunken phones (4 k ohms at DC) at 1 kHz by using a measuring bridge and obtained the quantities given in Fig. 2. REA is caused by the electro-acoustical transducing process. The AF source has to provide the real power for REA as well as, necessarily, for the DC coil resistance, and foremost the reactive power for the phone coils (2.5 H) that are to move the membranes. In order to obtain the maximum of power transfer the magnitude (amount) of the overall phones´ impedance ZAF (16 k ohms for my phones) must match the AF source. Again I preferred to experimentally find the optimum AF load: I connected in turn 14 phones and speakers of different impedance to the set, partly connecting two of them in series or parallel, which in total provided 20 load impedance values between 80 ohms and 75 k ohms in magnitude. From the AF voltage measured across these load impedances I determined the AF power. The coupling to the RF signal generator was readjusted to retain RF matching each time the AF load was changed. Fig. 6 shows the obtained results when using a) a diode 1SS16 at low RF power (1 µW) and b) with a 1N914 at higher power (1 mW). The optimum AF load impedance turned out to be, resp., 1.2 and 3 k ohms. In order to simplify matters the diode was in this experiment fixed to the 1/3 tap at the tuned circuit. This meant a compromise as regards matching and generally did not produce quite the maximum of achievable AF power. The dashed curves of higher AF power in Fig. 6 were obtained when the diode was connected to the top of the tuning coil (as discussed above).



(Fig. 6: measured AF power versus AF load impedance.)

In order to present my measurement results in a more general form, Fig. 7 shows the AF power obtained as a function of the AF impedance now divided by the respective occurring source (= diode) resistance. The curves indicate that the maximum is reached when ZAF has a value of 50 to 70 percent of the diode resistance. The simplifications introduced above, like choosing the fixed 1/3 tap, are probably the reason for not reaching a higher percentage. But we can say to be roughly correct with our predictions. - Sometimes it is suggested to match just available phones (speaker) to the diode by using a suitable transformer. I found this only helpful if the mismatch was extremely high. In the other cases the winding and iron losses of the transformer, as well as the inductive shunt, tend to

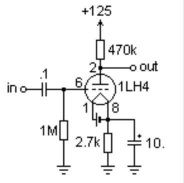


Pentode Voltage Amplifier

A pentode voltage amplifier with a gain of about -33 is shown at the right, using a 1U5, a sharp-cutoff pentode which comes in a 7-pin miniature package. Note the polarity of the filament supply, which is important, since it supplies a little grid bias as well. Since the voltages and currents are low, you can use the same 1/4W resistors used with transistors. The capacitors must have an appropriate voltage rating, of course. The 10M resistor is a "grid leak" that will prevent the grid from going positive. It charges on the positive excursions of the grid to provide whatever bias is necessary. The plate current was 61 µA, and the screen grid current was 20 µA. The screen grid potential is above the plate's in this circuit. Pentodes offer the capacity for controlling the plate current through the screen voltage, so that the plate voltage can be what is required. For higher voltages, the screen grid is normally at the same or lower potential as the plate, on the average. The grid voltage cannot be measured accurately with a DMM or scope, since the bias conditions will be disturbed. However, you will find that the amplifier works very well. Measure the gain the usual way, with a function generator and oscilloscope.

Radios for motor cars had 6 V DC available, which was right for 6.3 V heaters, but could not be used for the B supply--even now, 6 V is inconveniently low for transistors. The usual solution was a *vibrator* supply. Contacts on the magnetically-driven vibrator armature converted 6 V DC to alternating

The 1LH4 is a loktal battery triode, which is shown in a voltage amplifier at the left. A similar octal tube is the 1H5-GT, and either will do in this experiment (but the basing is different!). A diode plate is included, connected to pin 4, and using the negative side of the filament as its cathode. The diode is not used in this circuit. The maximum plate voltage is 110V, but this can be exceeded a little in testing. I measured the characteristics for plate voltages of 125V and less, at grid voltages of 0.0, -0.5 and -1.0V (relative to the negative end of the filament). The plate current was always less than 1 mA. My results were $\mu = 62$, $g_m = 340\ \mu S$, and $r_p = 182 k\Omega$. These values are close to the published values. The characteristics are noticeably curved in this region, so the parameters will vary with plate current. The apparently low value of transconductance is quite reasonable for the small plate currents at which the tube is used.

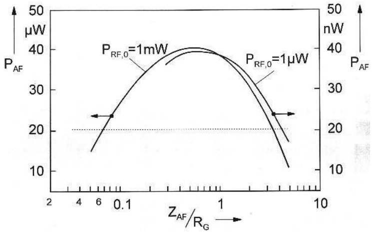


Voltage Amplifier

The amplifier shown gave a voltage gain of G - -40, with an input of 0.4V peak-to-peak. The plate current was about 100 µA, which made the plate voltage 77V and the grid bias -0.26V. The bandwidth of the amplifier was quite good, roughly from 10Hz to 50kHz. It should be noted that the total power drain of this circuit is no more than 83 mW, of which most is the filament power. This is very economical for a normal-sized tube. The tube remains quite cool in service, incidentally.

dissipate more AF power than is gained by providing the right transformation ratio. One also has to consider that the human ear cannot register small changes in acoustical power. Alterations like those shown above the dashed line in Fig. 7 will hardly be noticed by the ear. Thus, it seems that the exact value of the AF load impedance is not of paramount importance as regards noticeable output power. But the general principle holds that a number of small improvements in matching, each of which will not produce any audible effect for itself, might in sum indeed be noticed by the ear.

So it turns out as an interesting and important feature that a high-impedance AF load, which is associated with a high DC resistance, will produce a high diode resistance (= AF source resistance), and vice versa. This means that the circuit has a self-optimizing tendency towards the matched condition. Regarding RF selectivity of the set, as another important quantity, a high AF impedance - leading to a high diode resistance - is of advantage. But the influence of the AF impedance in this respect is not particularly pronounced. I measured an increase in -3db RF bandwidth by a factor of 2.5 when the AF load was decreased from 100 k ohms. At 720 kHz (local station) this bandwidth was 20 kHz in my set, which yields a total loaded Q factor of 36 - leaving room for improvement (see Secs. 2 & 3).



(Fig. 7: Measured AF power versus AF load impedance normalized to AF source resistance.)

7. Computer Simulation of Crystal-Set Behaviour

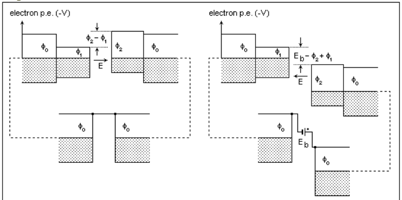
As a summarising investigation I simulated the overall performance of the circuit shown in Fig. 2, using the analysis program SPICE. An RF frequency of 1 MHz, an AF of 1 kHz, and a modulation factor of m=0.4 were assumed. The tap on the coil L2 was held fixed at 1/3 of the windings from earth. The I(V) equation given in Sec. 4 served to describe the diode behaviour.

because of space-charge effects. When you measure the V-I characteristic of a diode, there appears to be a small positive plate voltage when the measured voltage is zero, and a little current flows. The reason is that the electrons are emitted with some kinetic energy, and a negative space charge is created around the cathode. The actual emission to the anode comes from the minimum of this potential well, which is several volts negative, and so the plate may be slightly more positive than this minimum, in spite of contact potentials.

Battery Tubes

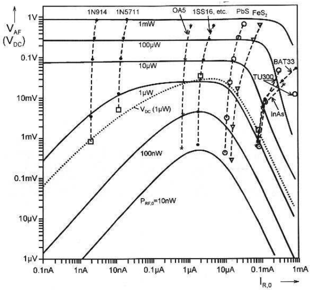
In the early days, nearly all radios (and other electronic apparatus) were powered by batteries, usually primary cells. With the rise of central power distribution, apparatus could be powered (and storage batteries charged) from the AC line, which was a practically unlimited, cheap source so far as radios were concerned. Transformers made high-voltage power supplies easy to build. Low-voltage filaments were largely replaced by 6.3V heaters and unipotential cathodes since power was now cheap. However, there still remained a demand for portable radios beyond the reach of power lines, and these were necessarily powered from batteries. The emphasis here had to be on small power drain, since battery power is expensive. Typical batteries were 1.5V and 3.0 V for filaments ("A" batteries), and 45V or 90V for plate supplies ("B" batteries), both the usual Leclanche "dry" cells. Filaments were generally used, because of their much greater efficiency in mA per watt. Typical battery tubes had filaments taking only 50 mA at 1.4V, only 70 mW. Plate currents were not high, since a few mA would be quite sufficient. For experiments with battery tubes, a D cell in a holder is an adequate filament supply, good for about 100 hours of service. Even an AA cell will do for experiments.

What happens in a vacuum tube is illustrated in the figure below. Suppose metal 1 is the cathode and metal 2 is the anode or grid, and the connections are made by a third metal 0 (there could be several such metals, but they all will act like metal 0). At the left, metals 1 and 2 are connected by a wire of metal 0. If metal 0 includes a DMM, it would show 0 V and, of course, no current would flow. However, the Fermi levels of all three metals must line up for equilibrium, making a level surface on which the electrons will not tend to roll one way or the other. It is clear that in this state of equilibrium, the surfaces of metals 1 and 2 will not be at the same electrostatic potential. An oxide-coated cathode has a work function of about 1.0 V, while a nickel plate or grid has a work function of about 5.0 V. When the Fermi levels align, the cathode is 4.0 V positive with respect to the other electrode.



Effects of Contact Potential

At the right, we have connected a source of emf E_b into conductor 0. A source of emf maintains a difference in the Fermi levels equal to its voltage. With the polarity shown, the cathode 1 has now become negative with respect to the anode 2, by an amount equal to the applied voltage less the difference in work functions. If the device is a thermionic diode, current can now flow from cathode to anode. The actual plate voltage is about 4 V less than the meter says, because of the contact potential. In practice, things are a bit more complicated,



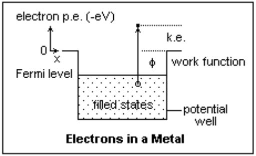
(Fig. 8: Simulated AF voltage across high impedance phones reverse saturation current of diode with available RF power as parameter (m=0.4). Measured values for various diodes included.)

The simulation results are presented in Fig. 8, which shows the AF voltage VAF AF obtained across phones of 16 k ohms AF impedance (4 k ohms DC) as a function of the diode reverse saturation current IR,O. Parameter is the available RF power at the tuned circuit, ranging from 10 nW (lowest sensible level; see Sec. 1) to 1 mW (about local-station level). Measured values pertaining to various diodes are entered for comparison. These measurement values lie in part slightly below, partly somewhat above the simulated curves, but agree in general. In practical diodes, particularly Schottky diodes, the effective reverse current increases noticeably with reverse voltage (i.e. with increased RF power). This had to be considered when inserting the measured values in Fig. 8, and accounts for the slight bending of the vertical lines to the right. The dotted curve in Fig. 8 shows the simulated DC voltage across the phones for 1 µW of RF power.

The simulated curves for VAF drop at the right side for diodes with high IR,O because of the adverse effect of reverse current as mentioned in Sec. 4. The decrease at the left of Fig. 8 for low RF powers results from the high turn-on voltage shown by diodes having a low IR,O. In these cases the associated smaller RF voltages increasingly fail to reach the turn-on of the diodes. The curves make clear how important it is to choose a diode which has the right value of reverse current. A definite maximum in sensitivity, especially pronounced at low RF powers, is found for diodes having a reverse saturation current of a few µA. Particularly the 1SS16 diode class is in this range, but also the OA5 and 1N34 perform not too badly, and good specimen of galena (PbS) crystals behave still satisfactorily. Hence this result is quite in agreement with what we already have found in Sec. 4. - When the AF voltages in Fig. 8 are used for calculating the AF power, the AF/RF detection efficiency can be worked out. It is found that the efficiency drops drastically for low RF levels, with one per mille being reached at 10 nW of RF. This agrees with the observations described in Sec. 1.

In the simulation the reverse breakdown voltage, at which in practical diodes the current starts to rise rapidly, was not included. For 1 mW, the highest RF power considered, the diode resistances have dropped to around 3 k ohms (Fig. 5a), calling for a low tap position on the coil. There the RF voltage is relatively low. My local station, with 3 mW of RF, produces 2.4 V (i.e. a peak-to-peak value of 6.7 V) at the required tap point, so that the negative peak only just reaches the breakdown voltage of -6 V for a 1SS16. In consequence, reverse breakdown seems not to be a particular limiting factor, even if the voltage across the tuned circuit might be somewhat

you measure with a DMM are not the actual voltages between electrodes, as seen by the free electrons, but differ by up to a few volts from the true potential differences.

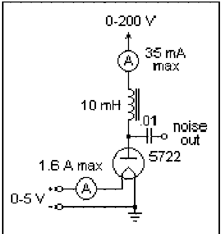


Electrons in a Metal

To understand this, let's review electrons in metals. One or more of the outer electrons of each metal atom is free to wander in the field of the ion cores, effectively binding the metal together. The positive ions create a potential well, as shown in the figure, and the electrons occupy the states of lowest energy. Note that electron energy increases upwards; since electrons are negatively charged, this means that voltage gets more negative upwards. There are not many more states than electrons, so they fill up the states up to some energy called the Fermi energy at 0 K. At ordinary temperatures, there is little difference, since the thermal energy is small compared to the kinetic energy of the electrons near the Fermi level.

If an electron is given enough energy, by thermal agitation or absorption of light, it may be separated completely from the metal and wander freely. The amount of energy necessary to just get outside is called the *work function*. Anything more goes into kinetic energy of the electron. The free electron can now be accelerated or decelerated by electric fields outside the metal in space. Work functions measured by thermionic, photoelectric and contact potential measurements agree roughly, but they depend sensitively on the surface preparation.

increases very rapidly beyond this point. The filament glows brilliantly, like an incandescent lamp, since its operating temperature is about 2400K, not the 900K of an oxide-coated filament. The filament current should not be allowed to exceed 1.6 A. If the power supply has current limiting, it can be useful here. By setting the plate voltage at near 200 V, you can see the saturation current as a function of filament current.



Testing the 5722

For two or more reasonable values of the saturation current, say 5 mA, 12 mA and 20 mA, record the current as a function of plate voltage and plot your results. For I_f = 1.5 A, the plate current saturated for about 50 V on the plate, approaching a value of about 12 mA. It is easy to find out what plate voltage to use to ensure saturation when making shot noise in this way. It is very difficult to make noise measurements in the usual breadboarding environment. I thought it just possible to have seen some on my 100 MHz scope with a plate current of 20 mA, without amplification. See the page on Noise for more discussion of noise measurements.

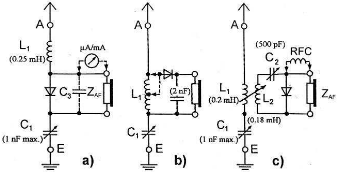
Contact Potential Effects

This is a good place to mention contact potentials, since you will probably run into their effects when experimenting with low-voltage tubes. Because of contact potentials, the voltages

higher in case there is a better Q factor. Assuming a constant available RF power, the RF voltage is proportional to the square root of the Q (resonance resistance). The low-barrier silicon Schottky diodes, which show a reverse breakdown in the range of -5 to -8 V, are thus well suited for use from the lowest to the highest levels of RF power generally occurring in crystal sets.

8. Some Remarks on Series-Tuned Circuits

Historically the first crystal sets, in the pre-broadcasting days, were of the kind shown in Fig. 9a. There the values of L1 and C2 pertain to my particular antenna/earth situation. For maximum power transfer in the circuit of Fig. 9a the combination of crystal plus load in parallel should match the impedance of the antenna source. In the latter the earth resistance represents the main resistive part which in the then primarily commercial stations had values of only 10 to 50 ohms. On the other hand, the crystal-diode resistances were around a few k ohms so that a considerable mismatch existed. For this reason one soon changed to the arrangement of Fig. 9b with diode and load now in series, and with the possibility to match the diode branch to the antenna by choosing the right tap on the coil L1. To increase the selectivity of the set a second tuned circuit was eventually introduced, as e.g. shown by Fig. 2.



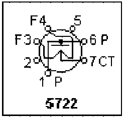
(Fig 9: Series-tuned circuits: a) Diode directly in tuned antenna circuit, b) Diode across tuning inductance (preferably tapped), c) Diode in separate series-tuned circuit, coupled to the series-tuned antenna/earth circuit.)

Using modern low turn-on diodes (1SS16 etc., Sec. 4) and having in general a higher earth resistance than in commercial stations, the circuit of Fig. 9a is however quite effective. Possibly paralleling of diodes is of advantage, depending on the actual source and load resistances. With two 1SS16 and employing a moving-coil speaker via a suitable transformer as the load I obtained an AF power of 180 µW from my local station. With ten 1SS16 and two moving coils of 16- ohms speakers in series as load, the obtained AF power of 210 µW approached the maximum possible after Sec. 1. Ideally the diode resistance should, in my case, about equal the 210 ohms of the antenna source (Fig. 1). Reverse diode current is not harmful, nor a possibly low reverse breakdown voltage. We have here a "current-controlled" case where voltages across the diode remain low with associated high currents, a few tens of a mV and some mA when I used the ten diodes. In contrast, voltage control is - more or less - experienced when parallel-tuned circuits are employed where high(er) voltages and low(er) currents exist. A set according to Fig. 9a, then, is a most simple hook-up for effectively receiving the nearest station. Substantially higher selectivity, approaching that of the set of Fig. 2, is offered by the arrangement with two series-tuned circuits shown in Fig. 9c. Since there any resistance (loss) in the circuit made up by L2 and C2 should be kept low for achieving a high Q factor, the diode(s) - preferably paralleled again - and the phones/speaker should be of low impedance. The RF choke might help to improve performance. The diodes found in the left of Fig. 4, particularly the backward diode TU300 (which is of little use on parallel-tuned circuits), operate excellently in the arrangement of Fig. 9c.

some surprise, that this correlated successive electrons so that they were emitted regularly to maintain a constant current, and therefore the shot effect was nearly completely eliminated. That is, a normal diode has no shot effect noise in its plate current.

The noise diode is designed so that at reasonable plate voltages, all electrons emitted by the filament are immediately drawn to the plate without forming much of a space charge. Since the electrons are emitted randomly, the anode current will show the full shot effect noise. This is done by purposely making the filament to have low emission. To do this, a tungsten filament is used. Noise diodes give us the opportunity to observe a tungsten filament, as well as temperature saturation.

An available noise diode is the 5722, whose basing is shown at the right. The 7-pin miniature tube was made as late as 1977, and now costs about $14, which is probably not much more than when it was new. The maximum plate voltage is given as 200 V, and the maximum plate current as 35 mA, so apparently the plate can dissipate 7 W. The plate has wings that make a good dissipation probable.



A circuit for testing the 5722 is shown at the left. Note that an RF choke is put in the plate lead to act as a load for the current fluctuations. This choke should be rated for the plate current employed. I connected a variable DC supply to the filament as shown, to pins 3 and 4, leaving the center tap alone. This supply should be rated at 2 A or more. Increase the filament voltage gradually, looking for the glow. There will be no plate current until the filament current reaches about 1.3 A, but it

As an example of the small signal diodes that are often combined with a triode or pentode in the same envelope, and share the same cathode, the 6AV6 or 6AT6 furnish good examples. The 6AV6 has its heater at pins 3-4, cathode at pin 2, and the signal diode plates at pins 5 and 6. The maximum current for each diode is 1 mA. I connected the two plates together for measurement, and took the current up to 3 mA, for which a plate voltage of 6.4 V was required. The curve of I against $V^{1.5}$ sagged a little at low currents, but the upper part was quite linear, showing a perveance of 0.085 $mA/V^{1.5}$ for one plate. The incremental resistance was 4.55kΩ, and V/I was 5.05kΩ at 1 mA. The current for one plate obeyed the formula $I = 0.15 + 0.085V^{1.5}$ mA. In this tube (and similar ones) the plates are flat, one on each side of the cathode.

The 1A3 seems to be the smallest signal diode of all. It was designed for portable measuring apparatus. The heater takes 0.15A at 1.4V (a D cell), connected to pins 1 and 7 of the 7-pin miniature envelope. The cathode is at pin 3, the anode at pins 2 and 6. The peak inverse voltage is 330V max., the maximum plate current 5 mA, and the average plate current 0.5 mA DC. Maximum heater-cathode potential is 140V. The anode is only a few millimeters high; most of the envelope contains only vacuum. The measured perveance was 0.075 $mA/V^{1.5}$.

The Noise Diode

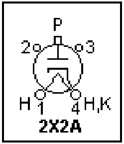
A special kind of diode should be mentioned here, because experiments with it are quite interesting. It is the *noise diode*, intended for the specific purpose of producing wide-band RF noise through the *shot effect*. Shot effect noise is fluctuations in the anode current due to the random collection of electrons. We have already mentioned that the anode current is controlled by the space charge around the filament. It was discovered, to

Modern 8/16- ohms headphones or, for stronger stations, directly the moving coil of a speaker are effective AF loads. - Backward diodes, being scarce these days, are tunnel diodes which have the typical current hump reduced to a flat region at about 200 µA height. The I(V) curve of the TU300 shown in Fig. 4 has in reality reversed polarity. For reasons of comparison with the other diodes the polarity was changed in the graph. Other BWD types are: AEY17 /29, 1N3539 /3543, TU1B.

In conclusion, the investigations sketched here have certainly enlarged my knowledge on crystal-set design, with the identification of the “best diode” and noticing the tendency of self-optimization which makes the set a sort of good-natured device. Other rewarding topics could not be covered, as there are, for example, more complex circuits for increased selectivity and for DX. Also short-wave crystal sets are fascinating since they provide DX from all over the world with simple designs.

Based on:
B. Bosch and M. Bussmann: Zur Empfindlichkeit von Kristallgleichrichtern und Halbleiter- dioden beim Detektorempfang.Funkgeschichte Nr. 93 (1993), pp. 275 - 285.
B. Bosch: Anpassungs- und Schaltungsfragen beim Detektorempfang.Funkgeschichte Nr.98 (1994), pp. 211 - 225.

Loktal equivalent to the types 80 or 5Y3 that are now much more expensive.



An excellent diode for observing the Langmuir-Child law is the 2X2A. This tube has a 4-pin base like the 82 phanotron discussed below, and the large, bell-like anode is brought out to a cap at the top of the ST envelope. The oxide-coated cathode thimble is easily seen. The heater takes 2.5V at 1.75A, so it can use the same transformer as the type 82. The rated DC current is 7.5 mA, and the maximum voltage is 4500V. A plate voltage of about 60V is needed to reach 7.5 mA plate current, so measurements can be made over a wide range of voltages. Plot your results as $I^{2/3}$ vs. V. A straight line will be found, that intercepts the V axis at -1.2V. The perveance of the 2X2 is found to be 0.0165 $mA/V^{3/2}$. The unusually low value is due to the large cathode-anode spacing.

The 6V3-A is a strange miniature tube with a cap on top that is the cathode connection. Its heater, connected to pins 4 and 5 of the 9-pin miniature base, takes 1.75 A at 6.3 V. The plate is connected to pins 2, 7 and 9. This tube is designed for the rugged service of a television *damper diode*. During horizontal retrace, the damper diode conducts, charging the boost capacitor while absorbing the large inductive kick. The peak inverse voltage is 6000 V, the peak current 800 mA, and the average current 135 mA. The large-diameter cathode tube and long plate imply a large perveance, which, in fact, is about 2.3 $mA/V^{1.5}$. This tube happens to be very cheap, but would serve as an excellent half-wave rectifier for practically any purpose. There are other damper diodes, such as the 6W4 and the 6AX4GT (perveance 1.42), that would have similar characteristics.

The 6H6 is an octal dual signal diode like the 6AL5, in a unique small metal envelope. The heater is connected to pins 2-7, the cathodes to 4 and 8, the plates to 3 and 5. 3 and 4 are one diode, 8 and 5 the other, and completely independent. It can be used for any reasonable service, such as AM detection, as a full-wave rectifier, or as a voltage doubler, so long as the current per plate is 8 mA or lower, and inverse voltages do not exceed 420 V. The voltage between heater and cathodes should not exceed 330 V. Measure the plate current as a function of the plate voltage up to 10 mA (the plate voltage will be about 7 V), and plot the current against the 3/2 power of the voltage. I obtained a rather straight line, showing agreement with Langmuir-Child, with a permeance of 0.5 $mA/V^{1.5}$. At 8 mA, the incremental resistance was 590Ω, and V/I = 785Ω. The 12H6 and 7H6 are similar tubes with different heater ratings and basing.

The 7Y4 is a typical small full-wave rectifier with an indirectly-heated cathode, like the more common 6X4 (miniature) and 6X5 (octal). This "Loktal" tube is inexpensive. Many of the common rectifier diodes are rather costly, for the curious reasons associated with the current tube market. The heater, taking 6.3V at 0.5A, is connected to pins 1-8 (as with all Loktal tubes). The cathode is pin 7, and the plates are pins 3 and 6. The peak inverse voltage is 1250 V, the peak current 180 mA, and the average dc current 70 mA. The heater-cathode voltage should not exceed 450 V. Measure the plate voltage for currents up to, say, 50 mA, and plot the results as for the 6H6. Again, we find a straight line and a perveance of 0.58 $mA/V^{1.5}$. Note that the plate voltage varies considerably as the current changes, from 4 V at 7 mA, to 16 V at 40 mA. Compare these voltages with those for a mercury-vapor *phanotron* as discussed in the next section. The 7Z4 is a somewhat larger full-wave rectifier (with perveance 0.40), the

Diode Characteristics

by Kenneth A. Kuhn

Oct. 3, 2007, rev. Sept. 3, 2009, draft –more to come

Introduction

This paper examines various electrical characteristics of a typical silicon junction diode. Useful mathematical relations are shown and illustrated with plots.

All of the plots are based on a typical sample of a very common small signal diode, the 1N4148 using a spreadsheet, diode_plots.xls, written by the author and posted on his web site: http://www.kennethkuhn.com . Because parameters vary from diode to diode, these plots should be interpreted as representative rather than absolute. Variations from diode to diode would typically be in the plus or minus several percent. To illustrate temperature effects, several of the plots are created using three temperatures: the high temperature is 100 C, the medium temperature is 25 C, and the low temperature is 0 C.

Forward current versus voltage

The standard equation for current through a diode is:

$$I = IS * (\exp(V/(n*k*T/q)) - 1) \quad \text{Eq. 1}$$

Where:
I is the current through the diode
IS is the reverse saturation current
V is the voltage across the diode (can be positive or negative)

n is a junction constant (typically around 2 for diodes, 1 for transistors)
k is Boltzmann’s constant, 1.38E-23 Joules/Kelvin
T is temperature in Kelvins
q is the magnitude of an electron charge, 1.609E-19 coulombs

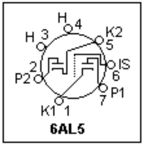
Looking at Equation 1 it would appear that the current should decrease as the temperature increases. The exact opposite is what really occurs. The reverse saturation current, IS, is a strong positive function of temperature as discussed below. The increase in IS with temperature more than offsets the effect of T in the exponential above.

The junction constant, n, is typically a constant at low currents and varies as the current becomes significant and may also vary somewhat with temperature. For this discussion, n will be taken as constant.

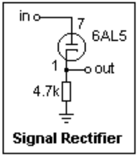
The sub-expression, kT/q, has units of voltage and is referred to as the thermal voltage and is typically around 26 millivolts at room temperature.

VT = k*T/q Eq. 2

Figure 1 shows a plot of Equation 1 for three temperatures. The strong temperature dependence can be clearly seen. There is a very important thing to note about the plot in Figure 1 –it is not possible to directly obtain the plot by applying a voltage and measuring the resulting current. Doing that will quickly destroy the diode because of a condition known as thermal run-away. Any power applied will heat the diode which in turn will result in the current increasing thus further heating the diode because of the negative temperature coefficient. This can be easily seen in Figure 1. The only way to collect the data for the plot is to apply a current and measure the resulting voltage



The 6AL5 dual diode, whose basing is shown at the right (7-pin miniature socket), is a typical signal diode. IS is an internal shield between the diodes. The two diodes and the shield are easily seen through the glass envelope, and you should notice how close the plates are to the cathodes. The close spacing means a large perveance, so only small plate voltages are required. Don't connect this tube directly across high voltages! A peak inverse voltage of 330 V can be resisted, and the DC plate current should not exceed 9 mA. Peak currents can go up to 45 mA if necessary, however. I measured the perveance as 2.42 mA/$V^{1.5}$, for one plate, a large value. The 6AL5 gives 9 mA with a plate voltage of only about 2.5 V! The heater, connected to H-H, pins 3 and 4, takes 0.3 A at 6.3 V.



Try the 6AL5 in the circuit shown at the left, which is a basic signal rectifier with a 4.7k load resistor. Feed it with the signal generator, and compare the output and input with the oscilloscope. Try input peak-to-peak voltages of only 2 V or so. You will notice that there is no "diode drop" with the 6AL5--it acts like a perfect diode, rectifying down to small voltages. We know how to do this with a semiconductor diode and an op-amp, but here it's done quite simply. The 6AL5 has an incremental resistance of only about 237 Ω, and is nearly linear. It is easy to run a plate voltage versus plate current curve with a low-voltage power supply. Keep the load resistor, and subtract the voltages at plate and cathode to find the plate voltage.

Thermionic diodes have now been completely superseded by semiconductor diodes, largely for economic reasons, physical size and the need for a filament supply. A silicon diode capable of carrying 1 A is available for \$0.04 or so, and takes up very little room. However, diodes can teach us a lot about thermionic emission and other interesting things. They do work rather well, and it is good to make their acquaintance.

The forward voltage (in the direction of current flow) of a diode is always relatively low, less than 15 V or so. The plate current is roughly proportional to the 3/2 power of the anode-cathode voltage (Langmuir-Child law), and the proportionality factor is called the *perveance*. The perveance depends on the geometry of the tube, increasing with larger area and closer spacing. It's remarkable that most diodes agree with Langmuir-Childs so well, in spite of different geometries. Since the voltages are low, contact potentials may affect your measurements. Contact potentials are discussed below in the section on low-voltage tubes. The easiest way to find the perveance is to plot $I^{2/3}$ against V, and to draw the best straight line. The intercept gives the value of the "true" zero plate voltage, and the slope, raised to the 3/2 power, is the perveance. Perveances range from 0.02 to 2.4 $mA/V^{3/2}$ for a representative assortment of 12 diodes of all types. There is no turn-on voltage drop for a thermionic diode, as there is for a silicon diode. Conduction begins immediately when the plate is positive with respect to the cathode, and stops immediately when the plate goes negative. It is easy to measure the V-I characteristic of a diode with a low-voltage DC supply, a voltmeter and an ammeter. I use a 100Ω resistor in series to make adjustment easier and safer. Thermionic diodes are not as easy to destroy as semiconductor diodes, and will take a good deal of abuse.

across the diode. Diodes are current operated devices and should never have a direct voltage applied from a low-impedance source.

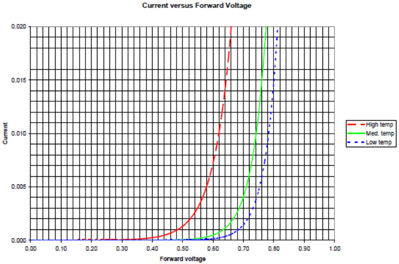


Figure 1: Forward Current versus Forward Voltage

The -1 term in Equation 1 can be omitted when the forward voltage is greater than about 0.1 volts at room temperature. A chart of the approximation and the exact is shown below.

For n = 1, VT = 0.026 volts

	Approximation	Exact	
V	exp(V/(n*VT))	[exp(V/(n*VT)) -1]	error%
0.00	1.00	0.00	infinite
0.02	2.16	1.16	86
0.05	6.84	5.84	17
0.10	46.8	45.8	2.2
0.15	320	319	0.3
0.20	2191	2190	0.05

Logarithmic forward voltage versus current

It is useful to plot the data in Figure 1 with the axis swapped as is done in Figure 2. An interesting observation in Figure 2 is that the forward voltage is a very linear function of the logarithm of the current through the diode. This characteristic is useful in building electronic logarithmic converters. It should also be observed that the slope is a function of temperature and that creates a complication as temperature compensation is required.

Equation 1 can be solved for the forward voltage:

V = n*VT*ln[(I / IS) + 1] Eq. 3

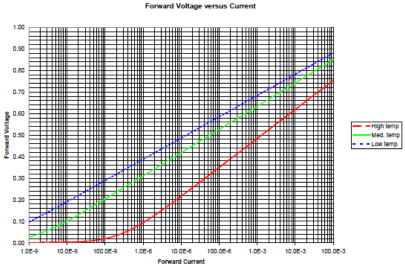


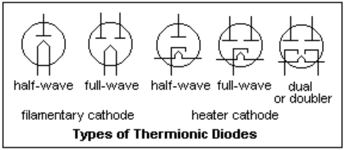
Figure 2: Forward Voltage versus Forward Current

Reverse saturation current

When negative voltages are applied to the diode the current becomes constant at –IS as the exponential term in Equation 1

Diodes

Thermionic diodes, like semiconductor diodes, are divided into *signal* diodes that handle small currents at low voltages, and *rectifier* diodes that handle large currents, often with large



inverse voltages. A diode has an electron-emitting cathode and an electron-receiving anode or plate. The arrangments of cathodes and plates in commercial tubes, and what they are called, are shown in the figure. Signal diodes are also often added to a triode or pentode, sharing the same cathode and with one or two plates. Current flows only from plate to cathode, and this unidirectional conduction is the purpose of a diode. Diodes cannot amplify.

Signal diodes always have indirectly-heated cathodes, so they are easy to use. It is only necessary to make sure that the heater-cathode voltage does not exceed specified limits, usually a few hundred volts. Rectifier diodes often have filamentary oxide-coated cathodes, since these cathodes are more efficient when large currents are needed, requiring less power. We are considering only vacuum diodes, *kenotrons*, in this section. Thermionic gas diodes, or *phanotrons*, will be treated below, since they have rather different properties.

Bakelite) to support the contact pins mechanically, taking the strain off the pressed-glass seal, which was made of lead glass. Around 1935, the metal envelope was developed, but there was still a base. The all-glass "miniature" tube was made possible by the "button seal" that supported the contact pins mechanically as well as bringing them through the glass, allowing the base to be eliminated and tube size to be reduced. The insides, or "cage," was the same size as in previous tubes, however. It is supported on its leads, which are welded to the contact pins before the envelope is fused in place and evacuated. The button seal is also used, in a larger form, on tubes designated by GB at the end of the type designation, and by Loktals. The final step in manufacture was "flashing" the getter, usually barium or magnesium, to perfect the vacuum by adsorption of any remaining gases, leaving a shiny coating. This was generally done by heating a loop inductively by RF from outside.

quickly approaches zero. That is why it is referred to as the reverse saturation current. The current is independent of applied voltage once a small voltage magnitude is exceeded. This current is very small and is typically in the low nanoampere region. The reverse saturation current is a strong function of temperature as illustrated in Figure 3. The following equation is a simplified model for the reverse saturation current and provides excellent results in normal operating regions.

IS = IK * exp-Eg/nVT Eq. 4

Where:
IS is the reverse saturation current
IK is a constant derived from n and IS at a known temperature
Eg is the bandgap voltage for silicon (ranges from about 1.20 to 1.28 volts)
n is the junction constant (typically around 2 for diodes, 1 for transistors)
VT is the thermal voltage as previously discussed

The reverse saturation current should not be confused with an imperfection in diodes known as leakage current from a high value shunt resistance across the diode jucntion. Leakage current is often many times larger than IS. Thus, IS can not be directly measured and must be computed using data from the forward bias region (see the section on measuring diode characteristics).

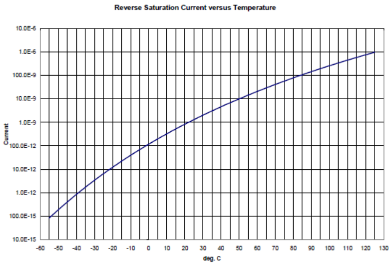


Figure 3: Reverse saturation current, IS, versus Temperature

Near Zero Bias

A theoretical plot of the diode current for the near zero bias condition is shown in Figure 4. Note that once the reverse voltage exceeds about -0.2 volts in magnitude that the reverse current is the value for IS and is independent of reverse voltage –this is why it is referred to as the reverse saturation current. Only the plot for room temperature (25 C) is visible. The low temperature plot is too small to observe on this scale except for the small forward bias region. The high temperature plot is very steep passing through zero volts. This illustrates why it is important to keep diodes cool when reverse current is an issue.

The plot in Figure 4 does not include the effect of leakage shunt resistance across the junction resulting from imperfections in the manufacturing process. This resistance varies from hundreds of millions to tens of billions of ohms. This shunt resistance results in an increasing current with applied reverse voltage and the current can become many

so the locking action guarantees that the tube will stay in the socket in spite of the small pins. The tubes are roughly the same size as an octal GT tube. Most are one size, but a few power tubes have a slightly longer envelope. There are no grid caps on any Loktal tube, and the heater connections are always to pins 1 and 8. Among the thoughtful features of loktal design, the type number appears in a hexagon on the top of the tube where it is visible from above, not on the side as on octal tubes. There is a dimple on the base corresponding to the key of the central pin, making the tube easy to orient for insertion. It seems that a lot of getter was used, so the tops of the envelope appear heavily silvered. The available types are only those used in AM and FM receivers. There are, nevertheless, enough types for a broad variety of experiments, and the prices are not excessive, so you may want to standardize on Loktals. Type numbers beginning with 7 have 6.3 V heaters, while type numbers beginning with 14 have 12.6 V heaters. There are some 7xx and 14xx tubes that are not Loktal, and some tubes that actually take a 7 V heater supply. One loktal rectifier, the 5AZ4 (a 5Y3 equivalent), has a 5 V filament. Loktal tubes designed specifically for battery-powered equipment had 1.4V filaments. The type numbers began with "1L." There were also rectifier and beam power loktals with 35, 50 and 70-volt heaters for AC/DC sets with series heater connections.

A tube designated simply 6N7 will be a metal-envelope octal tube with a 6.3V heater. A 6N7GT will have a cylindrical glass envelope. A 6N7G would have a shouldered glass envelope of the graceful shape designated ST. The electrical characteristics of such tubes were the same, whatever the envelope shape.

A very important part of vacuum-tube technology was bringing the metal leads through the glass envelope. Coefficients of expansion must be exactly matched, and the seal must be strong. Originally, tubes had bases (usually "duodecal" tubes were used in TV sets. Miniature tubes were not miniature, simply tubes with a button seal and all-glass envelope closely fitting a normal-sized cage. Subminiature tubes were actually miniature. Sometimes connections to grid, plate or (rarely) cathode were made to caps at the top of the tubes. In small tubes, these caps have a diameter of 1/4". The pins are numbered consectively clockwise, starting from the left of the index key for the octal, or to the left of the wider space, for the miniature, always looking at the bottom of the tube. This is shown for the octal base at the left. Pin numbers are given in the circuit schematics here. Most sockets have pin numbers marked. You will need to get sockets for the tubes you study, one for each type of socket. Solder wires to the tab at each pin that can be inserted in the solderless breadboard. I use the resistor color code for the pin numbers. A convenient octal socket fixture is available that comes with screw terminals for making connections. It was intended for relays, but is very useful for tube experiments. Heater connections for octal tubes are typically (not always!) to pins 2 and 7, and often to pins 3-4 on 7-pin, or 4-5 on 9-pin, miniature tubes. Sometimes halves of the heater can be connected in series or parallel, for two different voltages. Sockets were originally mounted in holes punched in aluminum chassis, secured by locking rings or by screws and nuts with a mounting plate. The chassis was, not surprisingly, the ground or common.

The "Loktal" tube was an excellent idea that was never universally adopted, mainly because miniature tubes took over in the 1950's. Since loktal was a trade name, RCA used "lock-in" instead, and you sometimes see "loctal." The loktal tube has an 8-pin button-seal (like the seal on miniature and octal GTB tubes). A natural metal base (of some aluminum alloy, apparently) shields the base of the tube and has a central pin with a circumferential locking groove. The pins project only 6 mm, and are 1.4 mm in diameter, much smaller than octal pins,

times IS. The effect of this resistance is most important at low temperatures where IS is very small. At high temperatures the effect of large IS swamps the resistance.

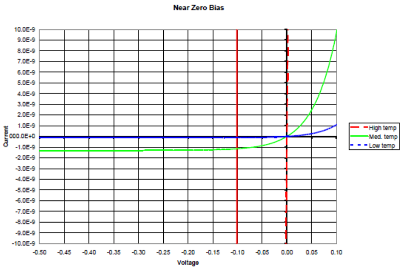


Figure 4: Near Zero Bias

(the vertical red line at -0.1 volts is a Microsoft bug and not part of the actual plot)

The plot in Figure 5 includes the effect of leakage shunt resistance. For this plot the resistance is 100 Mohms which was chosen for illustration and is very low compared to typical values of 1000 Mohms or more.

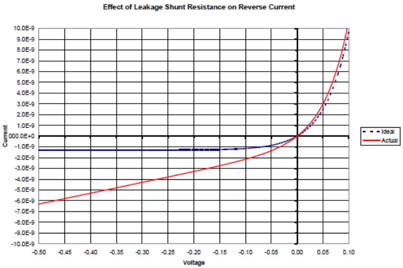
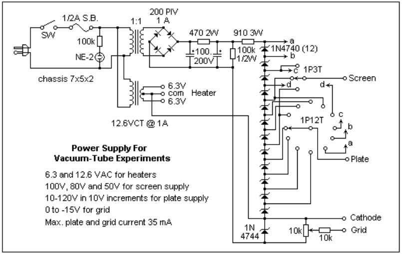


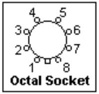
Figure 5: Effect of leakage shunt resistance

Temperature effect on forward voltage
With a constant current, the forward voltage drop of a diode has a very linear negative slope with temperature. The slope is also a function of current which is why constant current must be used. This characteristic is useful for building electronic thermometers. This characteristic is shown in Figure 6. Note that the slope becomes steeper as the current through the diode is reduced. For practical temperature measurements it is important that the diode not heat with the applied current. The power dissipation should be less than 1 milliwatt. This means that the constant current should be no more than about 1 milliampere. The plot in Figure 6 is based on Equation 3 but the x-axis is temperature instead.

main B+ supply. This cannot be done, of course, if the B+ voltage is adjusted using a variable transformer.

An all-in-one economical supply for vacuum-tube measurements is shown below. It uses an inexpensive isolation transformer from All Electronics, and can be made for about $30.00. The most expensive single part is the aluminum chassis. The grid potentiometer could be a precision 10-turn pot, but this would be expensive, and an ordinary carbon or plastic potentiometer (1/2 W or better) will be satisfactory. The maximum plate voltage of 120V and maximum plate current of 35 mA is adequate for many measurements. If you use a three-wire line cord, ground the chassis to the green wire. If you use only a two-wire line cord, it is probably better not to ground the chassis.





Vacuum tubes come with metal or glass envelopes, and in latter days with either the familiar octal arrangment of 8 pins, or as miniature glass tubes with 7 or 9 pins. There were earlier bases with four, five or six pins. Later, 12-pin miniature "compactron" or

The RC ripple filter is worth the expense. Waveforms are shown at the left. The waveform at node "a" is the familiar one for a "tank" capacitor, and the ripple is fairly large. Since the impedance of a 100 μF capacitor is only 26Ω at 60 Hz, the ripple is reduced by a factor of almost 25. At 300V output and a load of 12 mA, the ripple is less than 0.1V, a very satisfactory result. Note that all that is left in the ripple is the 60 Hz component. The filter would work even better on a full-wave rectifier, but here it is very satisfactory, better and more economical than larger capacitors. Of course, a filter choke could be used for an even better result and less voltage drop, but this would double the cost of the supply.

You will also need a heater transformer, which can be quite small if supplying only one tube that requires 0.3A. The transformer can be put in a box with an on-off switch and convenient terminals. Ground the heater supply (at a center tap if one is provided) to the B+ ground, to avoid excess voltages between heater and cathode. If you have a 12.6V CT (center-tapped) secondary, you can supply both 6.3 and 12.6 V heaters. Many 12.6V miniature tubes can also be connected for 6.3V. Tubes whose designations begin with "1" have filaments that can be supplied from a single D cell. Obtain a holder for the cell so connections are easy. 6.3 V was chosen to be compatible with 6 V car batteries, but the supply is usually AC. Many rectifier diodes use a 5 V filament or heater supply, apparently for historical reasons.

A "C" supply, for the grid bias in measuring characteristics, can be any isolated low-voltage supply of say, 15V, and a potentiometer can be used to pick off a variable voltage, since little current is involved. A separate high-voltage supply for screen grids may also be convenient, though it is easy to pick off the necessary voltages with a Zener or a VR tube from the

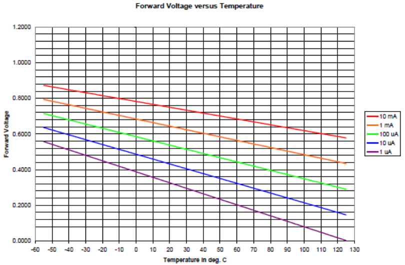


Figure 6: Forward Voltage versus Temperature

Temperature slope of forward voltage

The slope of the forward voltage versus temperature curve varies with the magnitude of the constant current as shown in Figure 7. This plot is made by taking the derivative of Equation 3 with respect to current.

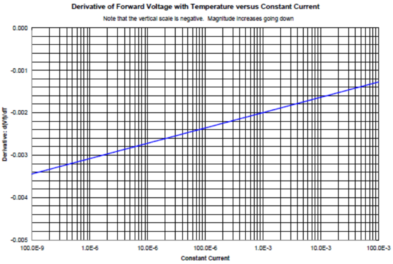


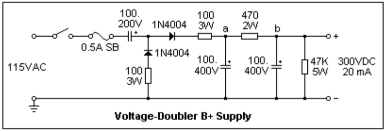
Figure 7: Derivative of Forward Voltage with Temperature versus Constant Current

Forward dynamic resistance

The operating point on Figure 1 represents a conductance (or its reciprocal function, resistance) for small signals (less than about 20 mVpp). The resistance is the reciprocal of the slope of the line through the operating point. This makes it possible to have an electrically variable resistor –the resistance is inversely proportional to the current. Electronically variable resistors are useful for building multipliers and gain control circuits. By taking the reciprocal of the derivative of Equation 1 (leaving off the -1 term) we have the forward dynamic resistance.

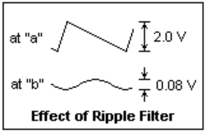
rd = n * VT / I Eq. 5

Observe that this resistance is a strong function of temperature. A plot of Equation 5 is shown in Figure 8.



Voltage-Doubler B+ Supply

An idea for an inexpensive B+ supply is shown at the right. The greatest expense is for the capacitors, which will cost about $15. It is based on a half-wave voltage doubler, and gives 300V for a 115V rms input. It cannot supply large currents, but is perfectly satisfactory for anything but power amplifiers. If supplied from a variable transformer, it becomes a variable supply for all voltages from 0 to 300. Note very carefully that one side of the supply is connected to the AC line, and this must be the grounded side, for your safety, and to avoid ground loops. You cannot ground the positive terminal of this supply to get a negative voltage supply (for use as a C supply, for instance). An isolation transformer, if you have one, would eliminate this hazard. If you don't have an isolation transformer, use a polarized plug to guarantee that the white wire is connected to the circuit ground. If you have a good ground, consider the old trick of connecting only one wire in the power cord, and using the ground to complete the circuit. It is best to observe the power ratings of the resistors and the voltage ratings of the capacitors. This circuit has been tested, except for the fuse. If the 0.5A slow-blow fuse fails, try a 1.0A. This fuse is to turn things off if a capacitor fails; nothing valuable is protected here, but it saves mess.



Effect of Ripple Filter

directly, because of the ground hazard. A variable transformer (Variac) is an *autotransformer* that does not isolate the output from the power line ground. I earnestly recommend that you do not work on AC circuits without isolating them from the service ground. The 110/220 adapters commonly available in 50W and 300W sizes, used for shavers and other small loads, should not be used, since they are autotransformers and do not provide isolation. They are, in fact, quite dangerous things, and should be used with great care. The supply was built in a 5" x 7" x 3" aluminum box, with an octal socket for the VR tube on top. The socket can be left vacant when the fixed voltage output is not required.

The voltage regulator requires a certain minimum current (about 5 mA) to function properly. If you are only drawing a few milliamperes from the supply, connect a 12k bleeder resistor across the output. Otherwise, the regulator will not adjust down to the lower voltages. Or, 220Ω and 10k fixed resistors, and a 15k pot, could be used at the voltage regulator, which would draw the necessary minimum current. The VR tube can be replaced by a high-voltage Zener diode.

A 25W isolation transformer is available at the date of writing from All Electronics (See the Your Laboratory page for a link) for $4.50. This transformer is surplus from the Power One firm, and is an excellent value. Solder a jumper between tabs 1 and 3, and another between tabs 2 and 4. The 120V input is connected between 1-3 and 2-4. The output tabs are marked B. This transformer would work well in the circuit above, or it could be put in a box and wired with line cord and output receptacle as a general isolation transformer. It should supply 200 mA without trouble, ample for our purposes.

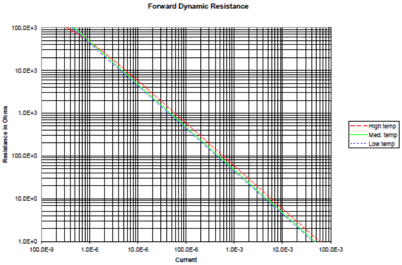


Figure 8: Forward Dynamic Resistance versus Current

Bulk Resistance

All diodes have a bulk resistance, Rb, that is in series with the diode. This resistance is typically very small (typically around 1 ohm or less) and its effects are insignificant at lower currents. At higher currents the effect is to reduce the voltage across the junction by the current multiplied by the resistance. This results in the steep theoretical exponential current versus voltage curve transforming into a linear slope. Equation 6 is Equation 1 with the effect of Rb included. Equation 6 (which omits the -1 term as that is not applicable for the higher diode currents) does not have a closed form solution as there is no known way to manipulate the equation such that either I or V is not involved with either a logarithmic or exponential term. Numerical solutions are the only way to solve Equation 6. Newton's method was used to generate Figure 9.

$$I = IS * \exp((V - I*Rb)/(n*k*T/q)) \quad \text{Eq. 6}$$

Figure 9 is a plot of Equation 6 and shows the effect of a 1 ohm bulk resistance for the room temperature case of the diode. Notice how the current is significantly less than theoretical for higher diode voltages. Note that the slope is nearly linear.

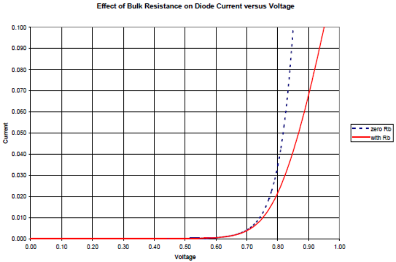


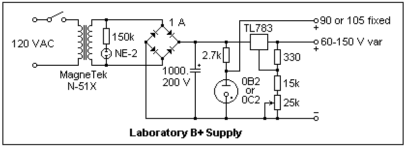
Figure 9: Effect of bulk series resistance

Sensitivity to light

The PN junction will respond to light –particularly longer wavelengths such as red and infrared. This feature can be a problem for diodes in transparent cases such as the 1N4148. The effect of light is to significantly increase the reverse conduction. A voltage can appear across the diode in response to light in high impedance circuits. This can be observed by connecting a diode (in a transparent case such as a 1N4148) to a DVM on the 2 volt scale and varying the applied light. In situations where light sensitivity could be a problem it is common to enclose a transparent diode in an opaque sleeve.

precautions and normal care, you will be fine. The solderless breadboard, DMM and oscilloscope can handle these voltages quite well. All the usual resistors and potentiometers are not afraid of 150V, so long as power ratings are observed. 1W and 1/2W resistors may be required in some places. Capacitors must be able to stand the voltages across them; many of those used with transistors will not be adequate. Keep a separate kit of capacitors rated at 100 V and above for this work. High-voltage capacitors are not needed everywhere, only in the plate circuits and for coupling from plate circuits.

The circuit of the laboratory B+ supply that I use for vacuum tubes is shown at the right, and the supply itself is shown in the



photograph below. It provides a regulated variable 60-150 V output, and a regulated fixed voltage output (for screen supply) created by a VR tube. VR tubes can be exchanged for different voltages. The MagneTek N-51X 115 V-115 V isolation transformer is available from Antique Electronics (see references), and a cheaper one from All Electronics. The transformer secondary has a DC resistance of 22Ω, which limits the surge current satisfactorily without having to add a series resistor. By no means eliminate the isolation transformer and use the 120 V household supply

were manufactured for voltages of 75, 90, 105 and 150 that were used like Zener diodes, handling from 5 to 40 mA. There is more information on glow discharges in Relaxation Oscillators, and on Zeners in Voltage Regulators. VR tubes are treated here because of their association with vacuum tubes, and the higher voltages involved.

Experiments

Vacuum tubes generally operate at higher voltages than transistor circuits. Like transistors, vacuum tubes are happier at higher voltages, which for receiving-type tubes, typically would be, say, 200 to 250V. It was once quite common to make DC power supplies for such voltages, using a transformer with a center-tapped secondary (say 200-0-200V), and a rectifier with double anodes and a common cathode, feeding a filter consisting of capacitors of 8 or 16 μF, and a choke of 10 H or so. It was not convenient to make a bridge rectifier with vacuum diodes (three separate filament transformers are necessary), so full-wave rectification with a center-tapped secondary was usual. These days it is rather difficult (and expensive) to acquire all these things, with the possible exception of the capacitors, which are now available up to millifarads at voltages up to 450 V.

The voltages normally used with receiving tubes are not high enough to be really dangerous, though a shock will not improve your day. If you are eager for new experiences, I can save you the trouble of finding out by saying that a DC shock is kind of like a hammer blow, not the zap of an AC shock, and does not paralyze, as an AC shock does. Shocks are given by current, not simple contact, so a good and old rule is to work with one hand in your pocket around high voltage. Always turn things off before making any adjustments or changes, of course, and be neat. Avoid touching bare metal. With these

There is a special type of diode known as a photodiode which is specifically made for sensing light. The case is transparent and often includes a lens to focus the light on the junction for increased sensitivity. Technically, it is usually infrared emission between around 800 to 1100 nanometers rather than visible light that these diodes are sensitive to. The sensitivity in the visible spectrum is usually poor except for what is known as blueenhanced diodes. These diodes can be operated either in an unbiased voltage mode or a current mode using a fixed bias around 5 volts. The current mode is very linear over a wide range. The current is sensed by a virtual ground electrometer.

Measuring Diode Characteristics

The mathematical model for diode operation can be completed by knowing IS and n. Neither of these are directly measurable but both can be calculated from data obtained by measuring the diode current at two voltages with temperature constant. We start with the following equation that omits the minus one term since our measurements will be at voltage levels high enough so that term is insignificant. Thus we have:

$I = IS * \exp V/(n*VT)$ Eq. 6

Equation 6 can solved for V by

$V = n*VT*\ln(I/IS)$ Eq. 7

$VH = n*VT*[\ln(IH) - \ln(IS)]$ Eq. 8A
$VL = n*VT*[\ln(IL) - \ln(IS)]$ Eq. 8B

Where H refers to the high voltage and high current measurement point and L refers to the low voltage and low current measurement point. These two points should be at currents less than 1 mA so that junction heating is insignificant. Good points to use are nominally 500 uA for the high value and 5 uA for the low value.

Subtracting Eq. 8B from Eq. 8A gives

$$(VH - VL) = n*VT*(\ln(IH) - \ln(IL)) \quad \text{Eq. 9}$$

Eq. 9 can now be solved for n

$$n = \frac{VH - VL}{VT * [\ln(IH) - \ln(IL)]} \quad \text{Eq. 10}$$

As a reality check, n is typically between around 1.6 and 2.0 for diodes and should be very close to 1.0 for transistors. Suspect some error if n is less than 1.0 or greater than about 2.2. The most common type of errors are using an inaccurate temperature or that the high and low measurements were made at different temperatures.

We can now calculate IS for the particular temperature the data was measured at. That temperature, T, must be known and the corresponding value of VT used in the following equation.

$$IS = IL * \exp{-VL/(n*VT)} \quad \text{Eq. 11}$$

As a reality check, IS should be around 1 nanoampere give or take perhaps a factor of ten for diodes. For transistors, IS will be extremely small –picoamperes or less.

The proof that these two values are correct is that they produce results that agree with measured data over a wide range when the model is plotted. The use of only two points for the calculations makes it extremely important to keep

also acted as an electrostatic shield between control grid and plate, reducing the Miller capacitance to extremely small values, 0.003 pF in the 6SK7. If the screen and suppressor grid are connected to the plate, the pentode operates as a triode.

A very curious and ingenious kind of tube was the *electron-ray* tube, used on receivers to give a visual indication of the accuracy of tuning to a station. Don't confuse it with the *cathode-ray* tube that uses a guided electron beam for oscilloscopes and TV receivers. It showed a luminous disk, with a dark sector. The dark sector was made as small as possible to achieve accurate tuning. It worked from the AGC (automatic gain control) voltage of the receiver. This is a feedback signal that tries to keep the signal amplitude constant at the output of the intermediate frequency amplifiers, increasing the gain for weak signals and decreasing it for strong. It is usually a negative voltage produced by rectifying the IF output. The tube has a thermionic cathode and a conical anode or *target* covered with cathodoluminescent phosphor (like a CRT), which glows from the 3 or 4 mA of plate current that flows when it is across 125V or more (up to 250). Control electrodes, of which there are two on opposite sides of the 6AF6, make two dark sectors that are widest when at 0V, and narrow as the control voltage approaches the target voltage. The control voltage is typically provided by the plate of a triode controlled by the AVC, such that full negative AVC cuts off the triode and makes the sector as small as possible. All this was cheaper and more graphic than a pointer meter.

Another kind of tube that we'll look at here is the glow-tube voltage regulator. The voltage across a glow discharge depends on the gas and the cathode material, and is almost idependent of the current through the discharge in the "normal glow" region, in which the glow does not completely cover the cathode, and expands to accommodate more current. Tubes toward the plate, to suppress the escape of secondary electrons. This is provided by a third grid, the *suppressor grid*, which is usually connected to the cathode. The tube with three grids: control, screen and suppressor, or grids 1, 2 and 3, is called a *pentode*, which turns out to be a superior voltage amplifier, fully equivalent to a transistor. A typical small pentode, the 6SJ7, has a plate resistance of over a megohm, and a transconductance of 1.6 mS.

An ingenious modification of the pentode has electrodes that shape and concentrate the electron beam instead of a suppressor grid, the negative space charge of the electrons doing the same work. These are called *beam power* tubes, and were good for power work, as the name indicates. A typical example, the 6L6, had a transconductance as high as 6.0 mS, and the smaller 6V6 about 4.0 mS. Both types were widely used for high-fidelity audio amplifiers, and tube amplifiers still have proponents. The same tubes were used in small amateur radio transmitters, which shows the versatility of vacuum tubes.

Receiving pentodes were also classified as *sharp cutoff* or *remote cutoff*, an example of designing tubes to fit their applications. A remote cutoff pentode had a grid with variable spacing, so that areas of wider spacing let electrons through when the grid was made more negative, when areas of smaller spacing were cut off. This effectively reduced the transconductance of the tube, decreasing its gain in response to increased negative grid bias, which was used for AGC (automatic gain control) in IF amplifiers. The 6SK7 was a very popular remote cutoff pentode used as an RF and IF amplifier. The 6SJ7, on the other hand, was a *sharp-cutoff* pentode, used as an audio voltage amplifier. The amplification factor has little significance with pentodes, as with transistors, and transconductance is the important parameter. The screen grid

measurement errors small. Much better accuracy is obtained by using a spread of data over a wide operating range (say currents from 10 microamperes to 1 milliampere) and then using linear regression to identify n and IS. That method is straightforward and is not shown here. For single situations one would enter data in a spreadsheet and then manually tweak n and IS in the spreadsheet model to minimize the sum of error^2 between measured diode voltage and the voltage computed from the model.

The model for the complete diode characteristic including the effect of temperature on IS can be determined by substituting the computed values for n and IS from above and the thermal voltage based on the measured temperature into the following equation.

IK = IS * exp(Eg/(n*VT)) Eq. 12

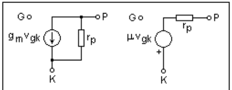
As an example, at 25 C the calculated values for a sample diode was 1.83 for n and 1.3 nanoamperes for IS. Eg was taken to be 1.23 –roughly a middle value. IK was then computed to be 338 amperes. These are the values used to create all the plots in this paper.

More to come …

shall examine these circuits in detail in the experiments, but this provides the background.

In order to provide higher voltage gain, the plate resistance must be reduced somehow. We recall that with a transistor, the analogous collector resistance was very large, and there was no problem with voltage gain. The plate resistance is the result of the effect of plate voltage on the space charge. This effect is not necessary for control, which is provided by the control grid, so what we need is to eliminate the effect of the plate voltage on the space charge. This is done by introducing another grid, the *screen grid* between the control grid and the plate. If this grid is held at a constant potential, the space charge is "screened" from the effects of changes in plate voltage. The screen grid is usually bypassed to ground by a capacitor, whose reactance at the lower corner frequency should be smaller than the resistance connecting the screen grid to B+. Some, but few, electrons are removed by the screen grid, since it is again a coil of fine wire. With this change, the plate characteristics become (nearly) horizontal lines, as for the transistor, and the plate resistance becomes large, approaching a megohm. The resulting tube is called a *screen-grid tetrode*.

Although tetrodes worked as expected, they had a serious defect. It happens that speedy electrons colliding with the plate knock out *secondary* electrons. In a triode, these are rapidly sucked back to the positive plate, and the same happened in a tetrode when the plate potential was higher than the screen grid potential. In normal operation however, especially with large voltage gain, the plate voltage has a large swing, and can become less positive than the screen grid. Now all these secondary electrons (and some of the primary ones, too) are attracted to the screen grid, and there is a definite sag in the characteristic in this region. To prevent this, it is necessary to establish an electric field at the plate that is always directed amplifiers, the allowable plate currents can be quite a bit larger, and hundreds of watts output is possible with relatively small tubes. The interelectrode capacitances of the 6J5 are on the order of 3.4 pF, and are significant at high frequencies.



Triode Equivalent Circuits

The small-signal equivalent circuits for the triode are shown at the right. The Norton source circuit is exactly the one for the FET. In the case of the triode, the plate resistance is always important, and cannot be neglected. The other circuit is just the Thévenin source corresponding to the Norton source. It shows the significance of the amplification factor, and is useful for triodes because of the rather small plate resistance. We did not find the Th%eacute;venin source for transistors very useful, and did not introduce a parameter analagous to the amplification factor for this reason.

It's easy to see that the maximum voltage gain achievable with a triode is μ, if the load resistance is much higher than the plate resistance. The rule we derived for the gain of a transistor amplifier as the ratio of the collector and emitter resistances holds here as well, expressed by $\mu = g_m r_p$. The quantity analogous to r_e is $1/g_m$, which is 333Ω for the 6J5. Mu is a rather modest number, so triodes are not good for high voltage gain. They make very good power amplifiers, however, since large currents can be controlled. Using the Thévenin source, the gain of a usual common-cathode amplifier (analogous to a common-emitter amplifier) is simply a voltage divider problem. There must be a plate resistor in series with the plate, or else the voltage would never change, but it should be as large as possible, and has only a small effect on the gain. We

Principles of Semiconductor Devices
http://ecee.colorado.edu/~bart/book/book/chapter4/ch4_4.htm
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Chapter 4: p-n Junctions
4.4.4. **I-V characteristics of real p-n diodes**

The forward biased I-V characteristics of real p-n diodes are further affected by high injection and the series resistance of the diode. To illustrate these effects while summarizing the current mechanisms discussed previously we consider the I-V characteristics of a silicon p+-n diode with Nd = 4 x 1014 cm-3, tp = 10 us, and up = 450 cm2/V-s. The I-V characteristics are plotted on a semi-logarithmic scale and four different regions can be distinguished as indicated on Figure 4.4.5. First, there is the ideal diode region where the current increases by one order of magnitude as the voltage is increased by 60 mV. This region is referred to as having an ideality factor, n, of one. This ideality factor is obtained by fitting a section of the curve to the following expression for the current:

$$J = J_s e^{V_a/nV_t} \qquad (4.4.39)$$

The ideality factor can also be obtained from the slope of the curve on a semi-logarithmic scale using:

$$n = \frac{\log(e)}{V_t\ \text{slope}} = \frac{\frac{1}{\text{slope}}}{59.6\ \text{mV/decade}} \qquad (4.4.40)$$

where the slope is in units of V/decade. To the left of the ideal diode region there is the region where the current is dominated by the trap-assisted recombination in the depletion region described in section 4.4.3.2. This part of the curve has an ideality factor of two. To the right of the ideal diode region, the current becomes limited by high injection effects and by the series resistance.

High injection occurs in a forward biased p-n diode when the injected minority carrier density exceeds the doping density. High injection will therefore occur first in the lowest doped region of the diode since that region has the highest minority carrier density.
Using equations (4.4.1) and (4.4.2), one finds that high injection occurs in a p+-n diode for the following applied voltage:

$$V_a = 2V_t \ln \frac{N_d}{n_i} \qquad (4.4.41)$$

or at Va = 0.55 V for the diode of Figure 4.4.5 as can be verified on the figure as the voltage where the ideality factor changes from one to two. For higher forward bias voltages, the current no longer increases exponentially with voltage. Instead, it increases linearly due to the series resistance of the diode. This series resistance can be due to the contact resistance between the metal and the semiconductor, due to the resistivity of the semiconductor or due to the series resistance of the connecting wires. This series resistance increases the external voltage, Va*, relative to the internal voltage, Va, considered so far.

$$V_a^* = V_a + IR_s \qquad (4.4.42)$$

Where I is the diode current and Rs is the series resistance.
These four regions can be observed in most p-n diodes although the high-injection region rarely occurs, as the series resistance tends to limit the current first.

varied, the plate current and voltage vary along the load line. The quiescent or operating point can be selected at some point along the DC load line, and so the DC grid bias can be found. This grid bias can be obtained from a C battery or equivalent, or from a cathode resistor, just as an emitter resistor is used with a transistor.

A cathode bias resistor is often bypassed by a capacitor if its negative feedback effect is not desired in dynamic operation. The reactance of the capacitor at the lower corner frequency should be equal to the resistance looking into the cathode (normally $1/g_m$ in parallel with the cathode resistor). The size of the cathode resistor has only a small effect on the size of the bypass capacitor. The capacitor never has a large voltage across it, and can be a low-voltage electrolytic. Only part of the cathode resistor can be bypassed if some feedback is desired.

A typical triode, of which the 6J5 is chosen here as the example, has a $\mu = 20$, $r_p = 6.7$k and $g_m = 3.0$ mS. Of course, the relation $\mu = g_p m_p$ always holds. The value of transconductance may seem small, compared to a transistor, but it should be remembered that it refers to a high-voltage plate circuit, and that the input impedance is infinite, so the power amplification is extremely large. The 6J5 is called a "medium-mu" triode. The similar 6SL7, a "high mu" triode, has $\mu = 70$, $r_p = 44$k and $g_m = 1.6$ mS. The 6SL7 is a *dual* triode, two independent valves sharing the same heater. The dual version of the 6J5 is the 6SN7.

The maximum plate voltage of the 6J5 is 300V, and the maximum plate current is 20 mA. Its maximum *plate dissipation* is 2.5W (product of average plate current and average plate voltage). This gives an idea of the ratings of receiving tubes used as voltage amplifiers. As power characteristics are curved, these quantities vary for different currents and voltages.

What we desire to represent is a function of two independent variables $I_p(V_p,V_g)$, which can be represented as a surface in three dimensions. Our various characteristic curves are orthogonal views along one or another of the axes. There are three such views possible, each directly related to one of the three parameters, of which we generally use only two, the ones mentioned here.

A vacuum tube carrying a current I with a plate voltage V dissipates power VI, just as if it were a resistor. However, the process is different. In a resistor, an electron gives up small amounts of energy to the lattice as it is accelerated and then is scattered. In a vacuum tube, the electron acquires a kinetic energy as it is accelerated, which it gives up all at once when it collides with the plate. It is not really correct to ascribe this to the "plate resistance," as some texts do, which is an incremental ratio. Since the plate is in a vacuum, the resulting heat can only be radiated or conducted down the supports. Much of the radiated heat is infrared, which is absorbed by the glass tube envelope. Note how plates are blackened to raise their emissivity and often provided with fins. Really large tubes had plates externally exposed capable of air or water cooling with elaborate seals to the glass parts. *Plate dissipation* is always a limiting factor in power applications.

The line marked *load line* shows the difference between the supply voltage V_{bb} and the voltage drop in a resistance R_L in series with the tube at any plate current, giving the plate voltage directly. The series resistance is mainly a plate resistor, but if there is a cathode resistor (for purposes of biasing) it should be included. There are generally different load lines for static (DC) and dynamic (AC) operation. As the grid voltage is

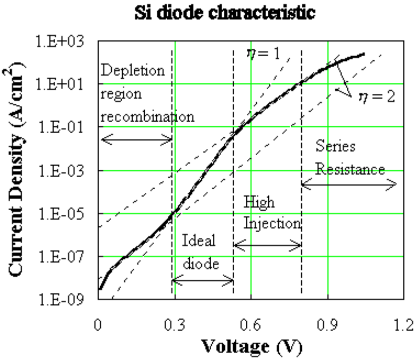


Figure 4.4.5: Current-Voltage characteristics of a silicon diode under forward bias.

characteristic, and a plot of I_b against V_g for a fixed value of V_p is called a *transfer characteristic*. From a family of either characteristics, the complete circuit behavior of the tube can be predicted. Unlike transistors, tubes of different types can have quite different (though similar) characteristics, so characteristic curves are much more important.

Let's begin with idealized plate characteristics for a triode, shown at the right. These are curved lines, but we represent them by straight lines for ease of understanding the various slopes and distances involved, which will be constant. Actual characteristics are not really far from straight lines, anyway. There is one curve for each grid voltage represented, which differ by a constant amount, here 2V. The horizontal distance between them, represented by the line labelled μ, is the amount the plate voltage must increase to hold the plate current constant; it represents the

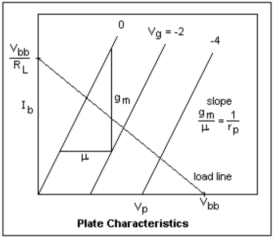
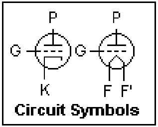


Plate Characteristics

relative influence of plate and grid on plate current, and is the *amplification factor*. The vertical distance between them shows how much the plate current changes for a change in grid voltage. The ratio is a conductance, called the *transconductance*, denoted by g_m, measured in siemens (mho). The slope of the curves is the ratio g_m/μ, called the *plate conductance* g_p. With vacuum tubes, the reciprocal r_p was always used, called the *plate resistance*. Since actual rectifier tubes, because they gave more current per watt of heating power. The indirectly heated, equipotential cathode that could be supplied by AC rather than by battery power was widely used after 1930.

A tube with just cathode and anode is called a *diode*, a term that has survived into the semiconductor age. Diodes were used for power or signal rectification, just like their semiconductor relatives. A "full-wave" diode has two anodes. When a control grid is provided, the tube is called a *triode*, and is used for amplification. Let's first study the peculiar circuit behavior of triodes, which will lead us to the reason for the addition of more grids, and the creation of the *pentode*, which turns out to act very much like a transistor.



Circuit symbols for a triode are shown at the right, and other tube symbols are derived from it. The connections are plate P, grid G, and cathode K or filament F, F'. These are analogous to the collector, base and emitter of a transistor, with the same polarities and direction of current flow as an NPN transistor. The circle is a part of the symbol. Grid connections can be to the right or left of the symbol, as convenient. A gas tube is indicated by a dot in the lower right-hand part of the circle. The heater of an indirectly heated cathode is usually not shown. A cold cathode (operating by positive-ion bombardment) is shown as a small circle. There are examples of these symbols below.

The important variables are the independent variables V_p and V_g, the plate and grid voltages (with respect to the cathode), and the dependent variable I_p, the plate current. A plot of I_p against V_p for a fixed value of V_g is called a *plate*

Germanium diode I-V Characteristics:
Ideal behavior analysis

Jose Maria Roman
https://www.uam.es/personal_pdi/ciencias/jmroman/presentations/jmroman_iv.pdf

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Abstract
This paper is devoted to the analysis of experimental results for a Germanium diode, and its comparison with the ideal diode theory. In particular, the I-V characteristics is measured for two different temperatures, showing a qualitative agreement with the predictions of the theory. Deviations from the ideal theory are presented and taken into account in a more detailed analysis of the forward bias characteristic. The temperature dependence of the reverse saturation current yields a value of the contact potential in good agreement with the known value of the gap for germanium.

1 Introduction

We analyze in this paper the experimental results for the I-V characteristics for a Germanium transistor ACY20, and compare them to the ideal diode theory. The ideal theory is based in simple assumptions of how the junction of two opposite doped semiconductors behaves when put together [1, 2, 3].

Measurements of the voltage and current across the diode at two different temperatures, t = 20dC and t = 80dC, are taken, showing a qualitative agreement with the predictions of the theory. Possible deviations from the ideal theory [1, 2, 3] are considered in a more detailed study of the forward bias characteristics, leading to several ranges of voltage where the ideal or ideal together with other mechanisms seem to dominate the current through the diode. The temperature dependence of the reverse saturation current allows us for obtaining an estimate of the contact potential, the potential barrier established between the two semiconductors of the junction. This value is in good agreement with the known value of the gap for germanium.

In section 2 we present the main features of semiconductors and junctions. Section 3 develops the simplest theory for the ideal diode, with a detailed calculation of the reverse saturation current from the underlying theory. Non-ideal mechanisms are outlined in section 4. The experimental set-up and measurement procedure are given in section 5, and section 6 contains the experimental results and their analysis. Finally our conclusion are presented in section 7.

2 The ideal diode

A diode is formed by the junction of two semiconductors doped with opposite sign, i.e., a p-n junction. The different electric character of the dopants at both sides of the junction establishes an electric field across the junction which determines the properties of the current which crosses the diode in the two different directions.

A n-type semiconductor is doped with elements called donors, because have electrons in excess of the intrinsic semiconductor elements (a semiconductor free of impurities or lattice defects is called intrinsic), e.g. phosphorous in silicon. The heavy doping density of donors, ND, determines the level of majority carriers, electrons in this case, so that nn0 = ND, while the

number indicates the heater voltage. 6 means 6.3V, 12 means 12.6V. Heaters are very forgiving of variations in voltage, but it is best to try to use the recommended voltages. AC is generally used, supplied from a small transformer. It is necessary to make sure that the difference in voltage between heater and cathode does not exceed 90V, so the heater AC supply should be grounded. It is usual to ground the center tap on the transformer for this purpose. Tubes for battery radios have plain filaments that are both heater and emitter, and must be supplied with DC. Their type numbers begin with "1" and are intended to be used with a standard dry cell of 1.5V. Larger tubes for high voltages have to use thoriated-tungsten filaments at 2000K,a bright yellow, to avoid damage from positive-ion bombardment, small as it may be. Heaters are sometimes called filaments, and the heater supply the filament supply, out of linguistic inertia.

Before the 1930's, each manufacturer used arbitrary designations for his tubes, and there was no uniformity or system. Tubes were first systematically identified in the U.S. by three-digit numbers, where the first digit denoted the manufacturer. For example, a type x10 was a power triode, a type x36 a screen-grid tetrode, where x was the manufacturer's number (usually omitted). Later, a new system was introduced where the first digit gave the filament or heater voltage, and the last digit gave the number of functional electrodes. This scheme was introduced by the RMA (Radio Manufacturer's Association) in 1934. A letter between these digits was assigned in order of introduction. For example, a 2A3 was a power triode with a 2.5 V filament (this popular tube is, remarkably, still in use for hi-fi amplifiers because of its low distortion!). This system was not comprehensive enough, and in the final system, the first number designated the heater voltage, but the remainder of the designation was arbitrary. Filaments were customarily used, especially in power and so electrons can pass through without hindrance. When it is made negative, it opposes the effect of the anode in creating an electric field, but does not attract any electrons, and so draws no current (except for the positive-ion current mentioned above). If it is made sufficiently negative, it can *cut off* the plate current entirely. If it is made positive, it can enhance the plate current, but then draws some *grid current* itself. The grid provides a sensitive control, using negligible power, of the large plate current, so the vacuum tube is a powerful amplifying device.

Early radio sets were battery-powered (domestic electrification was in its infancy, and absent in rural areas, when radio began), and a convention was established for identifying the batteries required. The filaments required low voltages at high currents (2W or more each), and their supply was called the A battery. The plates required high voltages at small currents, perhaps 90V at a few mA, and their supply was called the B battery. Grid bias was required, to hold the grids negative, demanding low voltages at small currents, and the corresponding battery was the C battery. The notations A and C were later little-used (except for actual battery radios), but the plate supppy came to be generally known as B+, and the letter B appears in subscripts of quantities referring to the plate circuit.

The cathodes of receiving tubes consist of a sleeve of nickel alloy coated with a compound of alkaline-earth oxides (Ba and Sr, usually). Inside is a tungsten heater wire insulated from the cathode with BeO or alundum (aluminum oxide) ceramic insulation. These cathodes must be heated only to about 850K (a dull red) to emit electrons in the amount necessary. Most receiving tubes require 6.3V or 12.6V for their heaters, at about 0.30A or 0.15A, respectively. Every tube type is identified by a type number, such as 6J5, where the first

density of minority carriers is determined from the mass action formula

$$np = n_i^2 = N_V N_C e^{-E_g/kT} \quad (2.1)$$

which leads to

$$n_{n0} = N_D, \qquad p_{n0} = \frac{n_i^2}{N_D} = \frac{N_V N_C}{N_D}\, e^{-E_g/kT}$$

(2.2)

where NV and NC are the density of states at the top of the valence and at the bottom of the conduction band respectively, and Eg is the band gap of the semiconductor.

Equally a p-type semiconductor is doped with acceptors, which have less electrons than the main element of the material (boron in silicon), and then trap one electron leaving a vacancy in the valance band, which is represented as a hole. Again, the doping density of acceptors, NA, determines the hole concentration in a heavily doped material, and the electron concentration is obtained from the mass action formula:

$$p_{p0} = N_A, \qquad n_{p0} = \frac{n_i^2}{N_A} = \frac{N_V N_C}{N_A}\, e^{-E_g/kT}$$

(2.3)

Now that we know the main features of n-type and p-type semiconductors isolated we put them next to each other. We assume here that we can treat electrons and holes independently. In this case the excess of electrons in the n-type side will diffuse into the p-type side. This diffusion effect produces a depletion of carriers around the interphase, known as the depletion region, the depletion will be supposed to be complete. Therefore producing an uncompensated charge to both sides of the junction which gives rise to an electric field (assumed to be contained exclusively within the depletion region). This electric field, in turn, generates a drift current opposite to the diffusion current, until an equilibrium between the two processes is reached.

$$I_{\text{diff}} = qD_p\frac{dp}{dx},$$
$$I_{\text{drift}} = qp\mu_p E, \qquad E = -\frac{dV}{dx} \quad (2.4)$$

Taking into account the Einstein relation Dp = kT¹p=q it is straightforward to obtain a relationship between the carrier density to both sides of the junction,

$$\frac{p_{p0}}{p_{n0}} = \frac{n_{n0}}{n_{p0}} = e^{qV_0/kT} \quad (2.5)$$

where V0 = Vn ¡ Vp is the potential drop across the junction generating the drift electric field, also known as built-in or contact potential. By using eqs. (2.2), (2.3) and (2.1) into (2.4) we can write

$$qV_0 = kT\log\frac{N_A N_D}{n_i^2} = E_g + kT\log\frac{N_A N_D}{N_V N_C} \quad (2.6)$$

which relates linearly the contact potential and the gap up to an additive factor depending on temperature, doping and density of states.

3 I-V Characteristic for an ideal diode

We can determine the general shape of the current flow through the junction considering a general argument for the mechanisms implied in the formation of the diode (see for example ref. [1], page 169).

The electrons emitted by the thermionic cathode form a negative *space charge* cloud around the cathode, dense enough that if no electrons are removed by attraction to the anode, the rate of emission is equal to the rate of return. When the anode is made positive, some of the electrons are attracted to it out of the space-charge cloud, and a *thermionic current* results. The amount of this current is given by I = A $V^{3/2}$, where V is the voltage from anode to cathode. This is called the Langmuir-Child law, and shows that electric field at the space charge produced by the anode controls the electron current. The cathode emits electrons copiously, so much that there are always enough electrons available to satisfy Langmuir-Child. Of course, at a sufficiently high anode voltage, the current may *saturate*, when all the emitted electrons are attracted to the anode, but this never occurs in normal operation, so small variations in cathode temperature have no effect. The current in a vacuum tube is said to be *space-charge controlled*.

If enough gas is present in the tube, the positive ions can counterbalance the negative electron space charge, robbing the anode of control and greatly increasing the current. Also, the positive ions can take over a large part of the electron emission at the cathode. Such tubes make efficient rectifiers, and the gas pressure can be quite low, as in some rectifiers, or rather high, as in a mercury-arc rectifier. In receiving tubes, positive ion collisions can destroy the delicate, high-efficiency cathode surface. Positive ions also cause small currents to negative electrodes that otherwise might be expected to carry no current at all. For all these reasons, receiving tubes have a high vacuum.

The electric field at the space charge that controls the current does not have to be created by the anode alone. A third electrode, the *grid*, is placed between the cathode and the anode, closer to the cathode. It is made of a spiral of fine wire, filaments used low voltages and high currents, and had a short life because of the high temperatures required for adequate emission. Most radio sets had a rheostat to adjust the filament current properly. Wehnelt emitters would have been quickly destroyed by positive-ion bombardment in these tubes. Apparently by accident, thoriated tungsten wire was used in a trial at the GE factory in Harrison, NJ in 1920 on a UV201 tube. Thoriated tungsten gave 75 mA/W of filament power, while tungsten gave only 1.75. Thoriated-tungsten filaments became popular for receiving tubes, such as the UV201A, which was an improved UV201, around 1924. Its filament required 0.25 A at 5 V, while the UV201's had required 1.0 A. Since tubes were now all *hard*, or high-vacuum tubes, indirectly heated oxide-coated cathodes, which gave copious emission at low temperatures, were used almost exclusively in receiving tubes after 1930. Not only did these cathodes have a long life, but were also *equipotential*, making circuit design simpler. Thoriated tungsten remained for transmitting tubes, where a rugged emitter was necessary because of the higher plate voltages, but even here tungsten was the only suitable choice for really high voltages, to avoid damage from positive-ion bombardment.

The rate of emission of electrons from a heated metal is given by the Richardson-Dushman equation, $i = AT^2e^{-b/T}$ A/cm^2, where T is the absolute temperature in K, and A and b are constants typical of the emitter. For tungsten, A = 60 and b = 52,400K, while average values for an oxide cathode are A=0.01, b = 11,600. The exponential factor has by far the largest influence, so emission increases rapidly with temperature. This makes thermionic cathodes very suitable even for heavy currents. In all tubes, electrons are emitted in far greater numbers than required; most simply return to the cathode.

As it was shown before there are two mechanisms which give rise to the diode conducting properties. On one hand there is the diffusion of electrons from the n-type side into the p-type side. These electrons have to overcome the potential barrier (contact potential) established by the electric field of the junction. This is achieved by a pure thermal activation mechanism

$$n \sim e^{qV_0/kT} \quad (3.1)$$

and therefore the density of carriers overcoming the potential barrier depend on the height of the barrier (given by the contact potential V0), and the temperature.

This diffusion current is compensated by a drift current originated by the electric feld across the junction. Differently to the diffusion mechanism this drift current is not very much affected by the variations in the contact potential between the two sides of the junction, since the electric feld acts on the minority carriers, and is the small density of them which determines the flow rather than the electric field itself.

We can assume therefore that in equilibrium V = 0 the diffusion and drift currents are compensated, and are both given by I0. When we apply a forward bias on the diode V > 0 the potential barrier is reduced from V0 to V0 ¡ V , and therefore the diffusion current increases by a factor eqV=kT , while the drift current remains unchanged. In the case of applying a reverse bias V < 0 the diffusion current will be reduced by a factor e¡qjV j=kT , while again the drift current is unaffected.

From the previous argument we can establish the I-V characteristic for the diode

$$I = I_0 \left(e^{qV/kT} - 1\right) \quad (3.2)$$

where I0 represents the equilibrium diffusion and drift current for V = 0. Notice that for and infinity reverse bias the diffusion current goes to zero and the only remaining one is the drift, such that I(V = ¡1) = ¡I0, a small finite value known as reverse saturation current. On the other hand, for large forward bias the current across the junction increases exponentially, and it is completely dominated by diffusion.

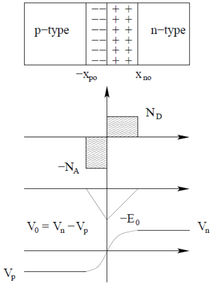


Figure 1. Ideal abrupt p-n junction, charge distribution, profile of the electric field and contact potential across the junction.

3.1 Reverse saturation current I0

Now that we have obtained the general expression for the I-V characteristics of the current across a p-n junction we will turn to compute the value of the reverse saturation current I0 under

"receiving" comes from their use in radio receivers, their principal commercial application, but refers to all small vacuum tubes for general electronic purposes. For the cathode-ray tube and making your own oscilloscope, see The Cathode-Ray Tube.

The electrons move from the cathode (K), the negative electrode, to the anode or *plate* (P), the positive electrode. Conventional current is in the opposite direction. The electrons are liberated at the cathode by heat--thermionic emission--or as a result of bombardment by positive ions, which can cause emission of electrons or even heat the cathode the required amount for thermionic emission. All receiving tubes employ thermionic emission, though we will note certain examples of *cold cathodes* in special cases. These were not usually really cold, but heated by ion bombardment rather than by a current supplied externally. The space in which the electrons move is not completely devoid of gas, so some gas molecules may be *ionized* by collision with speedy electrons, when an electron is knocked off, leaving a *positive ion*. The positive ions move in the opposite direction to the electrons (but their current is in the same direction, since they are of opposite charge). The effect of positive ions in a receiving tube is very small, because of the very high vacuum that is used.

Self-heated electron emitters are called *filaments*. The carbon filaments of the Edison Effect were soon replaced by metallic emitters, usually tantalum or tungsten, which were used by Fleming and de Forest. In Germany, Arthur Wehnelt discovered in 1903 that barium or calcium oxides baked on a platinum base emitted copiously, and used these *oxide-coated* emitters on vacuum rectifiers evolved from discharge tubes, which he patented in 1904. However, the use of *soft* tubes, which contained residual gas, demanded the use of rugged tungsten filaments, which dominated in 1910-20. These

1904 (B.P. 24850), where a filament and plate were arranged in the same envelope in a rather low vacuum, which could be used as a rectifier, or as a rather insensitive radio detector. In 1907, Lee de Forest patented the *triode* (which he called the Audion; the term "triode" was not used until much later, after it threated to become a trade name), in which a third electrode, the grid, was introduced to control the electron stream. This made a more sensitive detector, but the amplifying property was not used at first, and de Forest, who did not understand well what was going on, defended gassy tubes with their gas amplification. The introduction of high vacuum, as well as improved materials and processes, especially metal-to-glass seals, created a very useful amplifying device that allowed great developments in radio, telephony and sound reproduction. Schottky suggested a screen grid between the plate and control grid to make the electron tube useful at higher frequencies in 1919 (and actually made tubes with a second grid, but this was for space-charge control), but this was only realized by Hull and Williams in 1928 in radio receivers. The metal tube was introduced in 1935, but glass envelopes never disappeared and were constantly improved. The final pattern of electron tube was the "miniature" or all-glass type, which became the predominant receiving-type tube after about 1945. Transistors were invented in 1948, and in the next decade were improved to the point where they could take over most of the amplifying applications of electron tubes at much lower cost, and with greater reliability. Electron tubes remain in use as cathode-ray tubes, magnetrons, X-ray tubes, and for handling large powers. They were remarkable devices, using many sophisticated materials and processes, yet were widely available at low cost. We shall look here mainly at examples of *receiving tubes*, the smaller amplifying devices that have been completely replaced by semiconductors in current practice, but nevertheless will deepen our knowledge of electronics, while being fascinating to study. The name

the assumption that it is generated by low level (minority carrier) injection. In this approximation the majority carrier density is unaffected.

As shown in fig. 1 we suppose as a good approximation that we can treat electrons and holes independently, and that the drop of the contact potential is produced within the depletion region.

Assuming that the depletion region is completely depleted of carriers we have two regions with opposite charge, and densities equal to the doping, NA for the p-type side, and ND for the n-type one. This uniform charge produces a linearly spatially distributed electric field and a quadratically distributed contact potential, as show in fig. 1.

Under the previous conditions the current across the junction is established by the minority carrier injection on the other side of the junction. For example, holes will be injected

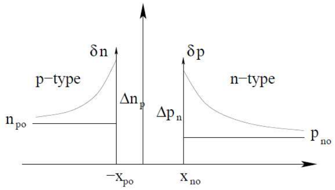


Figure 2. Profile of the low level injection in the p-n abrupt junction.

from the p-type side into the n-type side, and its concentration will decay to the equilibrium n-type hole concentration, pn0, due to recombination of the injected holes, as shown in fig. 2. Then, the concentration of holes in the n-type semiconductor is given by

$$\delta p = \Delta p_n \, e^{-(x-x_{n0})/L_p} \qquad (3.3)$$

and a similar expression for electrons,

$$\delta n = \Delta n_p \, e^{(x+x_{p0})/L_n} \qquad (3.4)$$

where xn0, xp0, Lp, Ln are the border of the depletion region in the n-type and p-type sides and the hole and electron diffusion lengths. ¢pn ´ p(xn0) ¡ pn0 is the excess of holes over the bulk n-type hole concentration at the n-type border of the depletion region. Similarly for ¢np ´ n(¡xp0) ¡ np0 for the p-type side.

The current passing through xn0 is given by the diffusion

$$I_p(x_{n0}) = -qAD_p \left. \frac{d(\delta p)}{dx} \right|_{x_{n0}} \qquad (3.5)$$

A similar expression to that in eq. (2.5) for the relation of carriers to both sides of the junciton under equilibrium conditions can be derived for the case of a forward bias, i.e. V > 0. The effect of a forward bias is to reduce the potential barrier to V0 ¡ V , and therefore the relationship between the holes in both sides of the junction is now given by

$$\frac{p(-x_{p0})}{p(x_{n0})} = e^{q(V_0-V)/kT} \qquad \longrightarrow$$

$$\Delta p_n \equiv p(x_{n0}) - p_{n0} = p_{n0}(e^{qV/kT} - 1) \qquad (3.6)$$

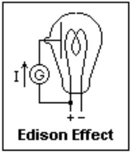
Theory of Vacuum Tubes

J. B. Calvert

Created 17 August 2001
Last revised 8 April 2003

http://mysite.du.edu/~etuttle/electron/elect27.htm#Diodes

Miniature vacuum tubes with cathodes of high-field-emitting carbon nanotubes are currently under study at Agere Systems in Murray Hill, NJ. A triode with amplification factor of 4 has been constructed, with an anode-cathode spacing of 220 μm, and a pentode is planned. Vacuum tubes may return to electronic technology! See *Physics Today*, July 2002, pp. 16-18.



Devices in which a stream of electrons is controlled by electric and magnetic fields have many applications in electronics. Because a vacuum must be provided in the form of an evacuated enclosure in which the electrons can move without collisions with gas molecules, these devices were called *vacuum tubes* or *electron tubes* in the US, and *thermionic valves* in Britain. In 1883, Thomas Edison observed that a current flowed between the filament of an incandescent lamp and a plate in the vacuum near it (see figure at the right), when the plate was connected to the positive end of the filament, but not when the plate was connected to the negative side (the plate was actually between the two legs of the filament). No important application was made of this unexplained *Edison Effect* at the time. In 1899, J. J. Thomson showed that the current was due to a stream of negatively-charged particles, electrons, that could be guided by electric and magnetic fields. Fleming patented the *diode* in

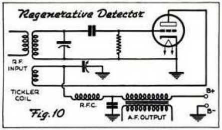


Fig. 10

ranged that part of the amplified signal is fed back to the grid to be reamplified. The feedback is supposed to be just a little too weak to sustain self-excited oscillation, the trick being to adjust the feedback just right.

A modification of this, the super-regenerative circuit, has several interesting and useful features. Though tricky, it's exceedingly sensitive and sharp-tuning. Essentially it is a regenerative detector, but a "quench circuit" consisting of a triode and an inductance-resistance network acts to interrupt its functioning at a supersonic frequency. The theory of its operation is ingenious. An ordinary regenerative detector with a little too much feedback will hesitate for a fraction of a second, then gradually sweep into oscillation. The super-regenerative detector has so much feedback as to cause oscillation, but the quench circuit keeps interrupting it before the oscillation can establish itself. This permits a tremendous amount of reamplification. The circuit is particularly useful on frequencies of 100 megacycles and above, where tube losses tend to be severe, for the principle permits B power to replace such losses.

The problem of demodulating F.M. signals is quite different from that with A.M.

In Fig. 11 the R.F. input is fed to a special intermediate-frequency transformer that has two secondary circuits, S_1 and S_2, each of which drives a conventional diode detector and filter circuit. (Usually one 6H6 dual-diode detector tube is used.) If the intermediate frequency of this receiver is 5.00 megacycles, the primary of the "discriminator transformer" would be tuned to the same value, S_1 tuned to 5.10, and S_2 to 4.90 megacycles. If the signal comes through at 5.00 megacycles, both secondaries and their associated detectors receive an equal voltage; and since the filter circuits are so arranged as to oppose the voltages, they cancel each other and no A.F. is produced. But when the signal frequency varies either up or down, it will approach the resonance of one secondary and move farther away from that of the other. One will get a decreased and the other an increased voltage, and a net voltage will appear across the audio output. Since the voltage across the audio-output terminals will increase as the R.F. signal departs from 5.00 megacycles, audio volume will be determined by the extent of frequency deviation. Audio pitch will likewise be determined by the number of times a second that frequency deviation occurs. The circuit therefore meets the required conditions of F.M. demodulation.

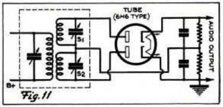
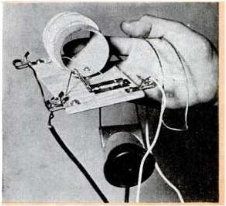
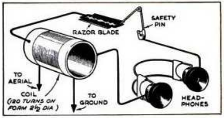


Fig. 11

Lt. M. L. Rupert's "foxhole radio," used on the Italian front, gets reception on a razor blade. O. B. Hanson, chief engineer of N.B.C., improved it by tying pencil lead to the pin point. Wire gauge is not critical, but good headphones, a long aerial, and patience in locating a sensitive spot on the razor blade are all requisite





Plugging (3.6) into (3.5) and differentiating we obtain an expression for the current at the border of the depletion region xn0

$$I_p(x_{n0}) = qA\frac{D_p p_{n0}}{L_p}\left(e^{qV/kT} - 1\right) \quad (3.7)$$

and therefore the total current, adding up the electronic part, is given by

$$I = qA\left[\frac{D_p p_{n0}}{L_p} + \frac{D_n n_{p0}}{L_n}\right]\left(e^{qV/kT} - 1\right) \quad (3.8)$$

from where we can easily read I0.

After using eqs. (2.2), (2.3) and (2.5) we can rewrite the reverse saturation current as

$$I_0 = qA\left[\frac{D_p}{N_D L_p} + \frac{D_n}{N_A L_n}\right] N_V N_C\, e^{-E_g/kT}$$
$$= qA\left[\frac{D_p N_A}{L_p} + \frac{D_n N_D}{L_n}\right] e^{-qV_0/kT} \quad (3.9)$$

which provide an explicit expotential dependence on the temperature and energy gap, or contact potential, assuming the terms inside the square brackets cancel each other or have a subdominant temperature dependence.

4 Non-ideal mechanisms

There are several aspects which have not been taken into account in the above derivation of the I-V characteristics, and further complicate the understanding of the diode behavior (ref. [1], page 211):

Recombination within the depletion region: We assumed that there was no generation-recombination processes in the depletion region. For material with a small density of carriers the depletion region can be quite large and then the probability of recombination inside the depletion region is higher even dominating the shape of the curve. The recombination rate is proportional to

$$R \sim 1/n_i \sim e^{qV/2kT} \quad (4.1)$$

Generation within the depletion region: For reverse bias what it is important is the rate of generation, since as soon as pairs are generated they are swept away by the enhanced electric field

$$G \sim W \sim \sqrt{V} \quad (4.2)$$

High level injection: As commented before we supposed that the density of majority carriers was not affected. However, as we increase the potential drop a large amount of carriers is taken from one side of the junction to the other, and injection for the majority carriers must be considered. When this is done we can see that the forward bias is dominated by the following form of the current:

$$I \sim e^{qV/2kT} \quad (4.3)$$

Ohmic losses: The potential drop inside the p and n regions, as well as that in the contacts, can be taken into account by introducing a series resistance Rs in the diode, which will influence the measurements as we use an external tension $V = V_{\text{ext}} - IR_s$.

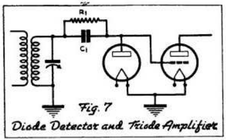


Fig. 7
Diode Detector and Triode Amplifier

The diode detector is commonly used today. It can handle very strong signals without overloading, has excellent fidelity, and is simple and reliable. However, it isn't extremely sensitive, since the filter system imposes a load that draws power from the tuned circuit. This in turn reduces selectivity.

Fig. 7 shows a modified diode detector directly coupled with a triode amplifier. Here the filter circuit operates somewhat differently; when a positive R.F. signal comes along, the 250-mmf. condenser C_1 transmits it freely, applying a positive potential to the diode plate and drawing electrons from the filament until the charge on the condenser is neutralized. This may take less than a ten-thousandth of a microsecond. When the positive peak declines to zero, the electrons are trapped on one plate of the condenser, on the plate of the diode, and on the triode grid. Their only escape is through the high-

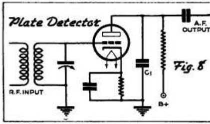


Fig. 8

resistance shunt, and it takes more nearly ten thousand microseconds than a ten thousandth of a microsecond for the charge to leak away. Thus the charge on the condenser can increase at radio frequencies but can decrease only at audio frequencies. Since it is applied directly to the triode grid, the triode plate current will follow A.F. modulation.

The grid of a triode is quite capable of collecting electrons from the filament, and consequently can serve as a diode plate in itself. In the grid-leak detector of years past, a single triode was used to perform this double function. The grid-leak detector was selective, though the fidelity was not high and it tended to overload on strong signals. It is seldom used today.

The circuit of a plate detector is shown in Fig. 8. The high value of the cathode resistance makes the grounded grid strongly negative with respect to the cathode, so much so as to cut off the plate current when there is no signal. When a negative signal reaches the grid, practically no change of plate current occurs, for it's already nearly zero. But when a positive signal arrives, the strong bias is decreased and plate current flows. This gives the nonlinear reaction necessary for demodulation. Condenser C_1 shunts the R.F. pulses to the ground. Able to handle fairly heavy signals, the plate detector shown gives good fidelity. It imposes no load on the tuned circuit and hence has good selectivity.

The infinite-impedance detector, so called because it has almost infinite resistance to R.F. between grid and ground, and hence places no load on the tuned circuit, is illustrated in Fig. 9. A positive signal applied to the grid permits the plate current to flow, and this causes a voltage drop through R_1 that drives the cathode positive. Since there is always some current flowing through the tube, the cathode is always somewhat positive with respect to the grid, and no grid current flows. Whatever the positive potential of the grid, the cathode will rise to match it, though it won't follow the negative swings as rapidly because of C_1. The result is that the potential across R_1 exactly follows the peaks of the R.F. signal. There is no amplification, but the detector has excellent fidelity, will handle strong signals, and is much used today.

Fig. 10 is the regenerative detector, of evil memory. It has excellent sensitivity and selectivity, but against these are poor fidelity, extreme unreliability, complexity of operation, and a bad tendency to turn into

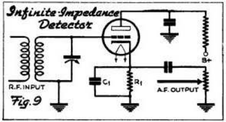


Fig. 9

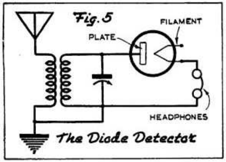
a transmitter able to send out squeals and howls into neighboring counties. Outlawed on broadcast frequencies now, it is still used by some amateurs on short-wave receivers. Basically it is a grid-leak detector so ar-

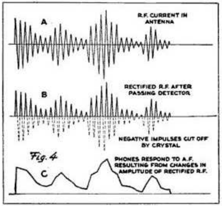
AND HOW THEY WORK

derstood phenomenon in which certain metallic oxides exhibit the ability of rectifying A.C. The plate-and-needle detector, essentially a metal needle resting lightly on a metal plate, embodies this principle, as does the razor-blade radio shown on page 209.

One of the most sensitive instruments known in the early days of wireless was the galvanometer. Unfortunately, it requires direct current; if it is fed A.C. of even moderate frequencies, the needle tries to swing one way on positive impulses and the other way on negative ones, with the result that it stands quite still. Before it could detect R.F., it had to be coupled with a rectifier that would intercept half the R.F. wave.

To improve the detector, the British Marconi Company called in Dr. J. A. Fleming, a leading English research physicist. Fleming had already done some notable research for a lamp-bulb manufacturer. The carbon-filament lamps of the time had a tendency to burn out at the positive end of the filament, due to the "Edison Effect." Edison had reported that a hot filament in an evacuated tube would retain positive but not negative charges—in modern terms, that it would give off electrons. Therefore electrons jumped from the negative end to the positive, carrying part of the current and heating the positive end by bombardment.

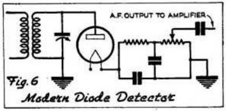




Dr. Fleming's studies of the Edison effect, plus his realization that the galvanometer would make a sensitive detector if it could be used with R.F. currents, gave him the answer—the Fleming Valve. It was a simple diode rectifier, a heated filament and a plate in an evacuated envelope, in series with a galvanometer and the antenna circuit. The diode permitted one half of the waves to kick the moving coil, but blocked the opposite set of half-waves. Before long headphones, taken over from the telephone, were used in place of the galvanometer and found unexpectedly efficient (Fig. 5).

Essentially, a demodulating detector is a nonlinear element—one that will pass more current in one direction than in the other—followed by a filter circuit. (A diode can be made to pass 95 percent of one half-wave and none of the other; a crystal may pass 80 percent of one and 40 percent of the other. So long as conductivity is dissimilar, demodulation will take place.)

In the simple diode detector shown in Fig. 5, the headphones themselves constitute the filter; the phone coils have sufficient inductance and capacitance to make an excellent R.F. trap. In Fig. 6, which shows a more modern version of the diode detector, a resistance-capacitance filter is used instead. The condensers and resistors are of such values that the condensers discharge through the resistors only at comparatively low frequencies—15,000 cycles a second or less—with the result that the ultrarapid R.F. pulses are completely smoothed out and only modulation remains. [Turn the page]



5 Experimental Procedure

In order to compare the theory of the ideal diode with the real diode behavior we will measure the current crossing a Germanium transistor ACY20 when a tension is applied across it, in forward and reverse bias.

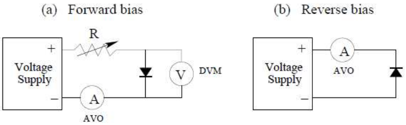


Figure 3. Experimental set-up for the forward and reverse bias configuration of the electric circuits to measure the I-V characteristic of the diode.

As shown in fig. 3a for forward bias an external voltage is applied to the diode and a series varistor, and the voltage drop on the diode, as well as the current, are measured. In fig. 3b the set-up for measuring reverse bias is shown. In this case the varistor is not necessary any more, since the variations in current as voltage is changed are small, and therefore the tension drop can be directly read from the voltage source.

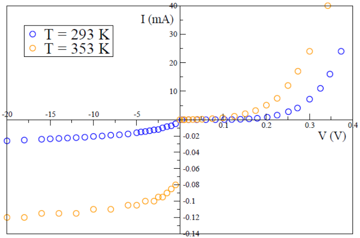
At room temperature we measure the I-V characteristics by controlling the current for forward bias ($I < 40mA$), and taking care of obtaining enough data for small values of the voltage ($0V < V < 0{:}2V$) in order to capture the expected exponential behavior of the characteristic. For reverse bias we can control the voltage itself and stop the reading before reaching the breakdown point.

In order to verify the exponential behavior of the reverse saturation current I0 with temperature (and energy gap Eg or contact potential V0), we heat the sample, while in the reverse bias set-up, for a given value of the applied voltage, V = ¡5V , from room temperature up to a temperature of about t = 80oC.

A new measurement of the I-V characteristic at about t = 80C is performed following the same procedure used for the room temperature case. We should obtain an increment of the current, in both forward and reverse bias, due to the thermal activation of the density of carriers, as predicted by the theory.

6 Experimental Results

We present in this section our results obtained from the previously described experimental set-up for a germanium diode. First of all, in fig. 4 we can see the I-V characteristic for forward and reverse bias, at room temperature (t = 20dC) and t = 80dC. (Notice the different scale used for the positive and negative axes of voltage and current).

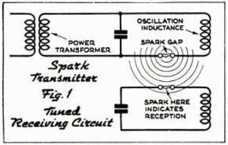


RADIO DETECTORS

By John W. Campbell, Jr.

IN RADIO parlance, the circuit in a receiver that strips incoming high-frequency waves of the audio impulses of speech and music they carry is usually referred to as the "detector" circuit. Actually the word "demodulator" would be more accurate, for a detector is any device that will reveal the presence of electro-magnetic radiation, whereas a demodulator separates the modulated signal from the carrier wave.

When Hertz sought by means of the first radio-frequency experiments to prove the existence of the electromagnetic waves indicated by the equations of Clerk Maxwell, he first had to devise a way of generating the radiations. The spark-gap oscillator answered that problem—he hoped. Hertz also needed a detector that would tell him if this theoretical electromagnetic energy was in fact present. The first detector was



simply a tuned circuit, matched to the transmitter. In their earliest form, Hertz's circuits were similar to those shown in Fig. 1; when powerful oscillations were set up in the transmitter, somewhat weaker oscillations were built up in the nearby detector circuit and a smaller arc leaped its spark gap.

Stimulated by Hertz's proof of the existence of electromagnetic waves, other scientists discovered new ways to detect them. Iron filings, it was found, tended to clump together in the presence of radio waves, and would then conduct a local current more readily. The result was the coherer (Fig. 2), which was in effect a crude amplifier, since it would turn on a local current when controlled by a distant signal. But it wasn't particularly sensitive, and the filings had to be "decohered" or loosened up, usually by a clapper device, before it could react to a new signal. It was thus suited only for code communication.

Next in order of simplicity after the coherer was the crystal detector, a circuit of which is shown in Fig. 3. The property of certain crystals, notably galena (sulphide of lead), of converting R. F. oscillations into uni-directional current was discovered early in this century. Though devised later than other, more effective detectors, crystals had the advantage of simplicity, low cost, and fidelity.

The operation of a crystal set is simple. The incoming R.F. (*A* in Fig. 4) is sent through a crystal which offers substantially more resistance to current flowing in one direction than in the other. The result, as indicated in *B*, Fig. 4, is a direct current, still pulsating at high frequency, with wave amplitudes that correspond to the audio modulation impressed on the R.F. When this current is applied to a pair of earphones, the diaphragms cannot follow the rapid R.F. pulses, but they can respond to the much lower frequency of the change in amplitude of the R.F.—that is, to audio modulations, shown at *C*.

Precisely why a galena crystal can offer differential resistances to A.C. has not yet been satisfactorily explained. Scientists believe that its action is related to barrier-layer rectification, another imperfectly un-

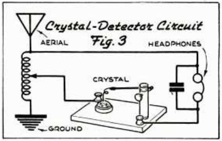


Figure 4. I-V Characteristics: values for t = 20dC and t = 80dC. Notice that the scale for positive and negative axis is different.

For forward bias we notice an exponential behavior (a more detailed study is presented in fig. 5) of the characteristics, and how the higher temperature characteristic allows for higher values of current in agreement with the assumption of a larger number of carriers available due to thermal activation.

In the reverse bias scenario we can see how the current tends to saturation. A saturation point is not reached probably because of generation current within the depletion region, or other kind of non-ideal mechanism as we will discuss later. Again the higher temperature curve yields a higher current.

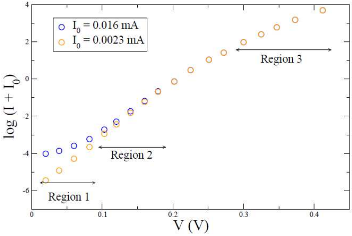


Figure 5. Ideality Factor: measurements in forward bias. Three different regions are identified, corresponding to different dominant mechanisms. Low voltage (region 1), low-intermediate voltage (region 2), and large voltage (region 3) regions are indicated in the figure.

The confirmation of the exponential behavior for the forward bias characteristic is presented in fig. 5. According to eq. (3.2), by plotting log(I + I0) against V we should obtain a straight line if the diode followed the ideal behavior. When non-ideal mechanisms are considered, as those presented in section 4, different regimes should be identified as the slope of the curve changes, since

$$\frac{qV}{kT} \longrightarrow \frac{qV}{AkT} \quad (6.1)$$

where A is known as the ideality factor. Usually A = 1 for the ideal mechanism, corresponding to low injection current, and A = 2 for high injection current (large voltage) or recombination current within the depletion region. In general intermediate values are expected since all the effects work at once.

There is a further uncertainty, since we do not know with precision the value of the reverse saturation current I0. Although in principle we are asked to take the value for a voltage of V = ¡5V , I0 = 0:016mA, we will argue that it is as reasonable to choose as saturation current the value for V = ¡0:05V , I0 = 0:0023mA. Other intermediate values of the saturation current would lead to intermediate results for ideality factors in the low voltage region. Curves for both saturation currents chosen are shown in fig. 5.

We can divide the points in fig. 5 in three regions, low voltage (0V < V < 0:1V , (region 1), low-intermediate voltage (0:1V < V < 0:2V , region 2), and large voltage (V > 0:3V , region 3) regions. By determining the slope of the linear fitting in each region we can obtain different ideality factors for each region and curve (m = q=AkT, see eq. (6.1)). The results for A are presented in table 1.

low band. About 1-2 wire diameter spacing between the two windings is recommended.

experienced when using a conventional inductor. If greater selectivity is needed at the high end of the BC band when using a conventional inductor, antenna coupling must be reduced and/or the diode must be tapped down on the tank to raise the loaded Q. Either approach results in a greater insertion power loss and a weaker or inaudible signal to the phones when tuning stations near the high end of the BC band . The low inductance (parallel connection) of the contra-wound inductor enables a 4 times reduction in bandwidth at 1710 kHz, compared to results with conventional inductor. This reduces the need to tap the diode down on the tank and re-match the antenna when one needs to increase selectivity, as mentioned above.

Note:
One could use two separate conventional non-coupled inductors, one of 250 uH and the other of 62.5 uH, instead of a contra-wound configuration. This is not recommended because the Q of the 62.5 uH inductor will probably be less than that of the 250 uH unit unless it is made physically as large as the contra-wound coil.and employs larger diameter wire. Also, when using the contra-wound approach the hot end of the inductor, when the two coils are connected in parallel, can be in the center of the overall unit, with the outer wire ends of the assembly placed at ground potential. This reduces electric field coupled losses from end mounting brackets and surroundings.

The inductances of the two connection configurations (parallel and series) of a contra-wound coil will depend upon how closely spaced the two windings are placed, but, the ratio of the inductance of the series to that of the parallel connection always remains at 4 times no matter how far or close together the windings are placed. Remember that overall distributed capacity is greater when using the parallel connection in the

Region	Io (mA)	A
1	0.016	3.10
2	0.016	1.45
3	0.016	2.58
1+2	0.0023	1.31

Table 1. Ideality factor for the different regions and curves in fig. 5

Our first curve (Io = 0:016mA) presents, in the low voltage region, a much larger value of A than expected from a recombination in the depletion region mechanism. In the low-intermediate region we could suggest a mix presence of recombination and ideal mechanisms. For the high voltage region a high value of A = 2:58 is obtained, which suggests that not only high level mechanisms are present. For slightly smaller values of the voltage drop (0:25V < V < 0:35V) a better agreement with high level injection current can be achived. The large value of A suggests that ohmic losses must be considered, which reduces the net current through the device.

We have considered as well the value I0 = 0:0023mA because at room temperature, kT » 25meV , the saturation of the current in reverse bias should be obtained for values of the reverse voltage as small as V » ¡0:05V . The existence of other effects in reverse bias (ref. [1], fig. 5-37, page 218), as pair generation in the depletion region, enhances the current, which will be misleading when trying to fit the ideal behavior. The smaller value of the saturation current in this case smoothers the curve for low voltages, and we can obtain a nice linear behavior spanning regions 1 and 2, with a value of the ideality factor A = 1:31, which is closer to ideallity. This value can be obtained when considering recombination in the depletion region due to interphase recombination, or defects in excess to what would be ideal.

Eq. (3.9) provides an explicit temperature dependence of the reverse saturation current with temperature, namely, an exponential behavior (provided we assume that the temperature dependence of the terms in square brackets cancels each other, or at least is subdominant). We can use the data of the saturation current, measured at V = ¡5V in the range of temperatures of t = 20d ¡ 80dC to obtain a value for the contact potential.

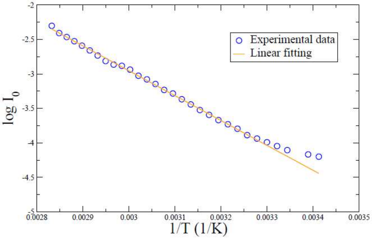


Figure 6. The contact potential is obtained from the slope of the linear fitting. This value is to be compared with the energy gap for Germanium. In the linear fitting the five most right experimental points where dropped.

In fig. 6 we plot the log I0 against 1=T , and from the slope of the linear fitting (m = 3607:9K) the value of the contact potential is V0 = 0:31V . This value of the contact potential is directly related to the value of the gap of the semiconductor

The value needed, using a contra-wound approach is 138.6 pF. One can derive, from data values in Figs. 1 - 4 in Article 28, that the Q of the common 365 pF, non-ceramic insulated variable capacitor (capacitor B), at 1710 kHz comes out as follows:

•If one uses a conventional 250 uH inductor tuned by 20 pF stray capacity with 14.65 pF more from the variable capacitor, the capacitor Q comes out at about 460.
•If one uses a contra-wound inductor that has 62.5 uH inductance with the two windings in parallel, tuned by 20 pF stray capacity with 118.6 pF more from variable capacitor B, the Q comes out at about 1770, 3.5 times as great! This translates directly to greater sensitivity and selectivity when using the commonly available 365 pF capacitor.

From Fig. 3 in Article #24 we can see that, at 1710 kHz, the Q of capacitor A, a ceramic-insulated, with silver plated plates capacitor manufactured by Radio Condenser Corporation, or its successor TRW, has a Q of 9800. This is much higher than that of capacitor B when using a conventional 250 uH inductor. Changing to a contra-wound coil while using the easily available capacitor B goes a long way toward a goal of reducing the effect of the variable capacitor on tank Q and loss at the high end of the band.

Less selectivity variation and less insertion power loss: Conventional inductor: The 3 dB down RF bandwidth will vary from 3.69 kHz at 520 kHz to 39.9 kHz at 1710 kHz, a variation of 11.6 times . Contra-wound inductor: The 3 dB down RF bandwidth will vary from 3.69 kHz at 520 kHz to 12.15 kHz at 943 kHz in the low band, and from 3.04 kHz at 943 kHz to 9.99 kHz at 1710 kHz in the high band, an overall variation of 4.00 times. This is about 1/4 of the variation

See 'The contra-wound tank inductor' in Part 3 of Article #26 and the paragraph after Figs. 2 and 3 in Article #29 for descriptions of two different contra-wound configurations.

Discussion:
Let us divide the BC band geometrically into two halves: This gives us 520-943 kHz for the low band and 943-1710 kHz as the high band. Assume, for ease of understanding, that the tank inductor for the conventional approach has an inductance of 250 uH.

Conventional 250 uH inductor: The whole BC band of 520-1710 kHz can be tuned by a capacitance varying from 374.7 to 34.65 pF.

Contra-wound 250/62.5 uH inductor: The low band of 520-943 kHz can be tuned, using the 250 uH series connection, by a capacitance varying from 374.7 to 113.94 pF. The high band of 943-1710 can be tuned, using the 62.5 pF parallel connection, by a capacitance varying from 455.76 to 138.60 pF.

For the purposes of this discussion, let us assume that antenna matching (see Part 2 of Article #22) is always adjusted to reflect a fixed shunt resistance of 230k ohms for driving the diode, over the full BC band. 230k ohms is also the RF input resistance of an ITT FO-215 germanium diode when fed a signal power well below its linear-to-square law crossover-point (see Article #10, points 1, 2 and 3 below Fig.1 in Article #15, Article 17A and Article #22). This setting approximates that for minimum insertion power loss (see Article #28).

Reduction of insertion power loss at the high end of the BC band (1720 kHz): The total tuning capacitance needed when tuning a conventional 250 uH inductor to 1710 kHz is 39.9 pF.

through eq. (2.6), which tells us that the value of the gap is given by the contact potential up to an additive term depending on the temperature, the doping, and the density of states. Therefore we can assume that the values of the gap and contact potential must be within the same order of magnitude, as it is the case since for Germanium Eg = 0:67eV .

As in the case of the ideality factor, here we also have to consider the uncertainty in the saturation current I0. The results for V = ¡5V could be affected by other effects which produce uncontroled deviations of the values of the contact potential and gap.

7 Conclusions

We have studied the behavior of a Germanium p-n junction, developing the simplest theory to explain it, namely, ideal behavior of an abrupt junction with low level injection current. We also outlined several corrections to this theory, which must be considered when analyzing the experimental data.

The I-V characteristic for the germanium diode was presented for room temperature (t = 20dC) and t = 80dC, showing the expected qualitative dependence with the applied potential, exponential increment with forward bias, and small values quite independent of the reverse bias.

A further study of the forward bias regime was presented to try to fit the different mechanisms involved in the transport of carriers through the junction. It was found that the low voltage region is consistent with the ideal behavior plus, possibly, generation within the depletion region. The high voltage region is consistent with a high level injection current mechanism and ohmic losses.

The dependence on temperature of the reverse saturation current I0 (V = ¡5V) yields a value for the contact potential generated by the junction, which is directly related to the energy gap of germanium. It was found that both values are within an order of magnitude, and the doping density and density of states account for the difference between them.

Acknowledgments

I would like to thank Rodrigo Ruiz Campo for his enthusiastic view of laboratory work, and very fruitful discussions. I also want to acknowledge discussions with Susana Fernandez Robledo.

References

[1] B. G. Streetman and S. Banerjee, Solid State Electronic Devices, Prentice Hall, fifth edition, 2000.

[2] S. M. Sze, Semiconductor Devices, Physics and Technology, John Wiley & Sons, 1985.

[3] J. P. McKelvey, Solid State and Semiconductor Physics, Harper & Row/John Weatherhill, 1966.

Ratio of effective resistance to d.c. resistance

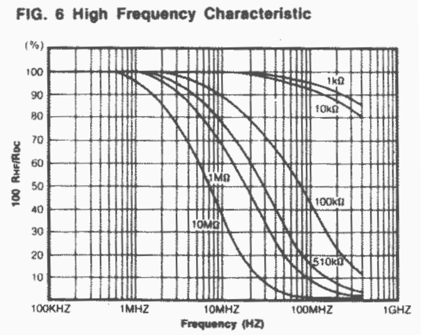
R (megohms d.c.) × f (Mc/s) =	0.1	0.5	1.0	5	10	20
I.R.C. type BTR (½W)	0.98	0.93	0.89	0.62	0.46	0.30
I.R.C. type BTA (1W)	0.95	.80	.71	.48	.37	.24
I.R.C. type BT-2 (2W)	0.80	.53	.40	.19	.14	.11
I.R.C. type BTS	1.00	.89	.79	.61	.57	—
I.R.C. type F (lower limit)	—	—	—	.84	.80	.75
Allen Bradley GB-1	0.85	.60	.48	.24	.17	.12
Allen Bradley EB½	0.90	.68	.57	.46	.23	.15
Speer SCT ½	0.92	.70	.60	.35	.27	.20

12. The effects from using the contra-wound dual-value inductor configuration in crystal sets as compared to using a conventionally wound inductor, both using capacitive tuning

Some quick facts:

- •Crystal sets using a conventional single-valued tank coil usually suffer from poor selectivity and sensitivity at the high end of the BC band.
- •Use of both connections of a contra-wound dual-value inductor enables the achievement of much higher selectivity and sensitivity at the high end of the BC band (series connection for the low half and parallel for the high half of the BC band).
- •There will be some small reduction in tank Q in the lower half of the BC band. One reason is that distributed capacity is greater in the series-connected contra-coil than in the conventional solenoid (the close-space adjacent ends of the contra-coil windings have 1/2 the tank voltage across them). Tank Q at the high end of the BC band is noticeably improved.
- •It is assumed that comparisons between conventional and contra-wound inductors use coils having the same physical dimensions and wire specifications. The inductance of the conventional solenoid is assumed to be about the same as that of the series-connected contra-coil.

A fact of interest that some may not know is this: The AC resistance of carbon composition resistors, and film resistors, to a much lesser degree, decrease with increasing frequency (the Boella Effect). This effect is strongest in high value resistors, above, say, 22k ohms and above 50 MHz (film resistors). The effect is noticeable in 500k and 1 meg units at lower frequencies. Low value resistors having short leads and resistances in the mid 10s to mid hundreds of ohms are quite free of this effect up through many hundreds of MHz. A typical graph of the ratio of AC-to-DC resistance vs frequency, of various values of conventional commercial axial-lead carbon film type resistors, taken from a Brell Components catalog is .



A chart providing similar info on carbon composition resistors, taken from the Radiotron Designer's Handbook, Fourth Edition, page 189 is .

DIODES
Dick Kleijer crystal-radio.eu
http://www.crystal-radio.eu/endiodes.htm

Which diode can I use best in my crystal receiver?

Maybe you think a diode with a voltage drop as low as possible, then also small signals at the detector circuit are detected. But diodes with a low voltage drop, also have a high reverse current (leaking current), this will load the detector circuit heavier, the Q of the detector circuit reduces, and with that also the voltage across the LC circuit. At a lower input voltage the diode will give much more losses, and it can happen that despite the lower voltage drop of the diode, you have less voltage at the load resistor. Besides that, reduction of circuit Q will also gives a less selective receiver.

For every 20 mV less voltage drop, the reverse current will approximately double.

Germanium, silicon, en schottky diodes. Depending on the material they are made from, we can distinguish germanium diodes, silicon diodes and schottky diodes. There are some more types, which are not discussed here.

Silicon diodes have the highest voltage drop (about 0.5 Volt) and are for this reason not very useable for crystal receivers. Unless we use a small DC bias current, which brings the diode already a little bit in conduction.

Germanium diodes have a low voltage drop (about 0.1 - 0.2 Volt) and are often used in crystal receivers. The properties like voltage drop and reverse current can vary a lot between two germanium diodes of the same type. In practice we can best test several germanium diodes in our receiver and then choose the best. The diode resistance RD of germanium diodes is most times rather low, and only useable in crystal receivers with a low Q (low sensitivity and low selectivity). For high performance receivers, we can better use a suitable schottky diode.

Schottky diodes have a voltage drop of about 0.25 Volt. The differences in properties between two diodes of the same type are often small. Schottky diodes with the correct resistance RD are very useable in high quality crystal receivers.

The given voltage drop is normally measured at a forward current of about 1 mA. Also if we measure the voltage drop of a diode with a multimeter, the test current shall be about 1 mA. But also below this voltage drop the diode can conduct current, and can rectify a RF (radio frequency) signal. Only the current through the diode is then much smaller. When receiving very weak stations, the current through the diode can be e.g. only 10 nA. At such a low current, the voltage drop of the diode is also much lower then at 1 mA.

Detected voltage as function of the input voltage. If we rectify a RF signal with a diode we can distinguish two situations.

Situation 1: Rectifying in the linear region

If the input voltage is high enough (well above the voltage drop of the diode at 1 mA), the output voltage of the diode will be about proportional to the input voltage.

So double input voltage, gives about double output voltage.

The output voltage is almost equal to the peak value of the input voltage.

The power losses in the diode are in this region very low compared to the rectified power.

The brand of resistor may be guessed by examining the smoothness and shininess of its surface finish, and looking at each end of the resistor to see where the wire exits. Allen Bradley resistors look the best. They have bright color code colors and a smooth shiny finish. At the wire exit point from the body one can usually see the appearance of a small shiny ring embedded in the plastic. Actually, this is part of the lead, shaped to be the contact electrode. Stackpole resistors look next best. They have somewhat duller colors on the color code and the surface is somewhat rougher and less shiny. The wires exit cleanly from the end of the resistor, no ring is visible. The Speer resistors have the dullest color code colors and a rougher surface than the Stackpole's. They usually look as if they have been wax impregnated. At the axial exit points from the body, a small copper colored dot may be seen next to the wire lead. This is actually the end of the lead, which was folded over and back on itself to form the electrode. The IRC so-called carbon comp. resistors can be identified by the visible 'mold-flash' marks on the body and ends. The colors are good, but the body is rough. Their end surfaces are slightly convex, not planar as in the case of the other resistors.

Remember, these resistors usually made spec. when new, passed incoming inspection and standard aging tests. Unfortunately, no aging tests could be made that covered the span of many decades.

It is interesting to note that the best resistors, from a long term resistance drift point of view turn out to be the AB units. They also cost the most. The Speer units cost the least and the Stackpole's were in between.

Ohmite carbon comp resistors I have seen looked like A-B units.

- •Audio transformers having too low a shunt inductance will reduce bass response. When using magnetic headphone elements, this can be partially compensated for by connecting the transformer to the headphones using a suitable capacitor.
- •Refer to Articles #2, #3,#5 and #14 for more info. Consider the 'Ulti-Match' by Steve Bringhurst at http://www.crystalradio.net/sound-powered/matching/index.html.

11. Long term resistance drift and frequency dependence of the AC resistance of low power resistors, etc:

From my early experience in the manufacturer of Blonder-Tongue products, the following is some insight relative to run-of -the-mill commercial carbon-composition resistors that we used:

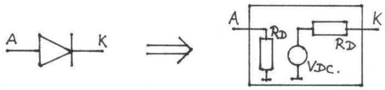
The process used by the resistor manufacturer is an important factor in the determination of long term resistance drift. Allen-Bradley (A-B) used their 'hot-mold' process, producing a more dense product then did the other manufacturers, as far as I know. The value of this carbon comp. resistor drifts the least, as a rule. Stackpole composition. resistors used their 'cold-mold' process and seem to drift more than do the A-B units. Composition carbon resistors mfg. by the Speer company, using their 'cold-mold' process drift more than the Stackpole resistors, as a rule. The IRC resistors that look like carbon comp. units actually are made by another process. They are called metallized resistors. My impression is that their drift is similar to the of Stackpole resistors. I have found that the IRC resistors usually generate much more low frequency noise when passing a DC current than the others. It seems, as a general rule, that the high value resistors drift more, over time, than the low value ones.

Situation 2: Recifying in the square law region

If the input voltage is low, lower then the voltage drop of the diode (at 1 mA) then the situation is completely different.
The input of the diode behaves for the RF signal like a resistor with value RD.
The output of the diode behaves like a DC voltage source in series with a resistor, the value of this resistor is also equal to RD.
The value of the DC voltage source is square law related to the amplitude of the RF input signal.
So double input voltage, gives 4 times as much detected DC voltage at the output
In the square law region the output voltage of the diode will be much lower then the input voltage, the diode gives much power loss between input and output.

The lower the input voltage, the higher the losses.
The higher the input voltage, the lower the diode losses.
When further increasing the diode input voltage, we gradually come into the linear detection region.
When receiving weak stations, detection takes place in the square law region.

Between the linear and square law region, there is a region not linear, and not square law but somewhere in between.
This region is not discussed here.



Equivalent circuit of a diode at low input voltages

Via this link you find a measurement on several schottky diodes, which shows detection in the square law region takes place at input voltages below 200 mVpp.

Diode resistance RD.

At zero voltage, diodes have a certain resistance.

This resistance at zero Volt we call RD.

The lower the reverse leaking current of the diode, the higher resistance RD.
When detecting small signals (in the square law region) the input of the diode also behaves like a resistor with value RD.

But how do we know the RD of a diode?

We can calculate it with the formula:
formula 1: RD = 0.000086171 x n x TK / Is
RD = diode resistance at zero Volt (unit: Ohm)
n = ideality factor, the lower this factor the better, between 1.0 and 1.1 is a very good value.
TK = temperature in Kelvin (= temperature in ºC + 273)
Is = saturation current (unit: A)
x = multiply

The values of n and Is can (sometimes) be found in the diode datasheet.

In the following table some types of schottky diodes, with the values for n, Is and Rd, the maximum reverse voltage and the diode capacitance at zero voltage.

Here are some practical experimental ways to vary the audio source resistance of a crystal radio set when receiving weak-to-medium-strength signals. A medium strength signal is defined as one at the crossover point between linear to square law operation (LSLCP). See the graphs in Article #15A.

- •Change the diode to one having a lower saturation current, such as from a germanium diode (1N34A) to one or several paralleled Schottky diodes such as the Agilent 5082-2835. Schottky diodes described as "zero bias detectors' have a high saturation current and are not suitable for most crystal radio set use. Schottky diodes described as "power rectifiers' usually have a high saturation current as well as a high junction capacitance. A high diode junction capacitance will reduce treble response. Too large a diode RF bypass capacitor in the crystal radio set can also reduce treble response. A side benefit from a change to a diode having a lower saturation current value, on some crystal radio sets is an increase in selectivity. This is because the RF load resistance presented by the diode to the tank is raised when the diode saturation current value is reduced. This reduced loading raises the tank Q and hence, increases selectivity.
- •Use an audio transformer between the detector output and the phones. A smaller step-down transformer impedance transformation ratio will raise the transformed diode source resistance seen by the phones. A larger ratio will decrease it.
- •If the headphone elements are in series, reconnecting them in parallel will reduce their impedance to 1/4 the previous value. This has the same effect as increasing the effective source resistance driving the headphones. If they are in parallel, series connecting them has the effect of decreasing the effective source resistance.

#9082 are suitable and are available from various distributors such as Alltronics, Digi-Key, etc.

Surface mount diodes manufactured using the SOT-23 package can be handled using Surfboard #6103. Diodes using the smaller SOT-323 package can be handled using Surfboard #330003. This includes many Agilent surface mount diodes useful in crystal radio sets. Packages containing multiple diodes exist that use the SOT-363 six lead package. They can be handled using Surfboard #330006. Agilent produces many of their Schottky diodes in dual, triple and quad form in the SOT-363 package.

It is recommended that anyone considering using Surfboards visit the above mentioned Website and read "Application Notes" and the "How-to Index".

10. How to modify the tone quality delivered by headphones: It is interesting to note that driving magnetic headphone elements with a high source resistance tends to improve the treble and reduce the bass response, compared to the response when the AC source resistance matches the effective impedance of the elements. Conversely, driving the headphones elements from a low resistance source tends to roll off the treble, and relatively speaking, improve the bass. With piezo ceramic or crystal elements, a high source resistance tends to reduce the treble and improve the bass response, compared to the response where the source resistance matches the effective impedance of the elements. A low source resistance tends to reduce the bass and emphasize the treble. Some piezo elements sound scratchy. This condition can be minimized by driving the elements from a lower resistance source.

type diode	n	Is at 25 ºC	RD at 25 ºC	maximum reverse voltage	capacitance	
5082-2835	1.08	22 nA	1260 kΩ	8 Volt	1 pF	datasheet
BAT85	?	?	± 200 kΩ ??	30 Volt	10 pF	datasheet
HSMS 2820	1.08	22 nA	1260 kΩ	15 Volt	1 pF	datasheet
HSMS 2850	1.06	3000 nA	9.07 kΩ	2 Volt	0.3 pF	datasheet
HSMS 2860	1.10	38 nA	743 kΩ	4 Volt	0.3 pF	datasheet

To decrease the value RD, we can connect more diodes in parallel, with two the same diodes parallel the value of RD shall halve.
With 3 diodes parallel, the value of RD shall be divided by 3 etc..

Diode resistance when using bias current.

We can decrease the value of RD by sending a small DC bias current (e.g. 0.1 uA) in forward direction through the diode.
The higher the bias current, the lower RD will be.

With the following formula we can calculate the diode resistance RD, when we make use of a DC bias current.

Formula 2: RD = 0.000086171 x n x TK /(Ib + Is)

RD= diode resistance at certain DC bias current Ib (unit: Ohm)
n= ideality factor of the diode
TK = Temperature in Kelvin (= temperature in ºC + 273)
Ib= DC bias current through the diode in A
Is =saturation current of the diode in A

A diode with a certain RD value at a certain bias current, gives the same receiving performance as a diode without bias current with the same RD.

Influence of temperature on "saturation current: Is"

The saturation current (Is-value) is strongly depending on temperature.
A temperature increase of 1 ºC will increase the Is value by about 7 %.
In datasheets, the Is value is most times given at 25 ºC.
If the diode temperature is not 25 ºC, but another value "T", then we must multiply Is with a factor 1.07^(T-25).
T = diode temperature
^ = raise to the power of

Ideality factor n

The ideality factor n of a diode indicates how good the diode performs with regard to an ideal diode.
A (not existing) ideal diode has a value of n=1.
At low input signals, the maximum available detected output power is proportional to 1/n.
So doubling the n will halve the output power (this only applies at weak signals).

signals, each of power P/2 and that there will be four output components, as stated above. They are:

- •The two second harmonic components (both of the same frequency and phase).
- •The sum frequency component (f1+f2) Hz, which will be of the same frequency and phase as the second harmonic components since f1=f2.
- •The difference frequency component (f1-f2) at a frequency of zero Hz.
- •If we filter the harmonic and sum components as well as the two original signals from the output, only the zero Hz signal will remain; and we call it the detected DC output.

A diode detector can be thought of as a "Black Box". If the DC output impedance of the detector is matched to its load resistor and the AC signal power source of P Watts 'available power' is impedance matched to the input AC impedance of a diode detector, the DC output power can closely approach the 'available power' from the AC source. This gives us another way to look at a detector. It can be considered to be a "Black Box" that changes incident AC power of frequency "f" Hz into output power of frequency zero Hz (DC). This is the detected DC output.

9. Using surface mount components in crystal radio sets:
A convenient way to connect to the tiny leads of small surface-mount diode and IC devices is to first solder them to a "Surfboard". Pigtail leads can them be soldered through holes drilled in the Surfboard conducting races for connection to a circuit.
A surface mount device such as the OPA-349 integrated circuit (Eight lead SOIC package) can be soldered to a surfboard such as that manufactured by Capital Advanced Technologies (http://www.capitaladvanced.com). Their Surfboards #9081 or

voltage (remember the equivalent voltage is 3 times the original source internal voltage). The 1/2 comes from the 2:1 voltage division between the resistance of the equivalent source of 90,000 Ohms and the detector input resistance of 90,000 Ohms. The ratio of the new detector voltage to the old is: 3 times 1/2 divided by 0.9 = 1.67 times. This equates to a 4.44 dB increase in power applied to the detector. If the input signal to the detector is so weak that the detector is operating in the square-law region, the audio output power will increase by 8.88 dB! This is about a doubling of volume.

7. Caution to observe when cutting the leads of a glass Agilent 5082-2835 Schottky diode (or any other glass diode): When it is necessary to cut the leads of a glass packaged diode close to the glass body, use a tool that gives a scissors type of cut. Diagonal cutters give a sudden physical shock to the diode that can damage its electrical performance. This physical shock is greater than one might expect because of the use of plated steel instead of more ductile copper wire. Steel is used, in part, because of its lower heat conductivity, to reduce the possibility of heat damage during soldering.

8. Several different ways to look at a diode detector: A diode detector can be thought of as a mixer, if one thinks of its input signal as consisting of two identical signals of equal power, in phase with each other. It is well known that if a common AM mixer is fed with two signals of frequencies f1 and f2 Hz, most of the output it generates will consist of the second harmonic of each signal and two more signals at other frequencies. One is at the sum frequency (f1+f2) Hz and one at the difference frequency (f1-f2). Additional mixer products can be generated, but they will be weaker than those mentioned and will be neglected in this discussion. In the case of an AM diode detector, we may consider that its input signal of power P Watts is in reality the sum of two equal in-phase

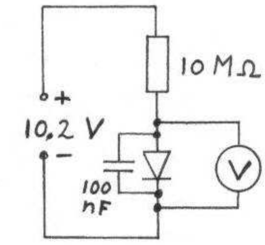
Diode capacitance

Between the two connections of the diode there will be a certain capacitance (capacitor value), when this capacitance is fairly high (e.g. 10 pF) the tuning range at high frequencies is limited.
At increasing reverse voltage across the diode the capacitance will reduce, also the detected voltage in a crystal receiver is such a reverse voltage.

Through this, the frequency of the circuit can shift upwards when receiving strong signals.
On the next page: experiments with a detector unit you find in table 3 a measurement about the frequency shift.

Measuring the Is value of a diode.

We can measure the Is value of a diode as follows:
Send a small current through the diode, the value of this current (ID) must be about 1 μA.
Measure the voltage across the diode (VD).



Circuit diagram for measuring the Is value of a diode.
The voltage across the diode is about 0.2 Volt.
The voltage across the resistor is about 10 Volts, so the current is about 1 μA.
The voltmeter must have a resistance of at least 10 MΩ.
The 100 nF capacitor reduces the influence of radio signals and hum on the measurement.

Calculate Is with the formula:

formula 3: Is = ID / (e^ (VD /(0.0257xn))-1)
Is = saturation current of the diode in nA
ID = current through the diode in nA, (1 μA = 1000 nA)
e = base of the natural logarithms, this is about 2.718
^ = raise to the power of
VD = voltage across the diode in Volt
n = ideality factor of the diode, if you don't know this value, take for instance: n= 1.08

An impedance-matched condition occurs when the resistance component of the input impedance of the crystal radio set equals the source resistance component of the impedance of the antenna-ground system. Also, the reactive (inductive or capacitive) component of the impedance of the antenna-ground system must see an opposite reactive (capacitive or inductive) impedance in order to be canceled out. In the impedance-matched condition, all of the maximum available power (See section on "Maximum Available Power" above) intercepted by the antenna-ground system is made available for use in the crystal radio set and none is reflected back towards the antenna to be lost.

Now we are at the point where confusion often exists: The voltage concept vs. the power concept. Let's assume that the diode detector has a RF input resistance of 90,000 Ohms. Assume that the antenna-loaded resonant resistance of the tuned circuit driving it is 10,000 Ohms. If one uses voltage concepts only, one might think that this represents a low loss condition. NOT SO! After all, 9/10 of the actual source voltage is actually applied to the detector. If one impedance matches the 10k ohm source RF resistance to the diode 90k ohm RF resistance via RF impedance step-up transformation (maybe connecting the antenna to a tap on the tuned circuit, and leaving the diode on the top), good things happen. (We will assume here that, in the impedance transformation to follow, the ratio of loaded-to-unloaded Q of the tuned circuits is not changed.) For an impedance match, the tuned circuit resonant resistance should be transformed up by 9 times. If this was done by a separate transformer (for ease of understanding) it would have a turns ratio of 1:3, stepping up the equivalent source voltage by 3 times and changing the equivalent source resistance to 90,000 Ohms. What now? Before matching, the diode got 9/10 of the source voltage applied to it. Now it gets 1/2 the new equivalent source and spreads out as it goes away from the antenna. One can prove that power is radiated by substituting a LED diode for the regular diode, getting physically close enough to the station and then tuning it in. The LED will light up (give off light power), showing that some power is being broadcast and that it can be picked up. Now back at home, if one tunes in the station one gets sound in the headphones. What activates one's hearing system is the power of the perceived sound. BTW, if one gets too much sound power in the ear for a long enough time, the power can be strong enough to break off some of the hair cells in the inner ear and reduce one's hearing sensitivity forever. The theoretical best one can do with a crystal radio set setup is the following: (1) Use an antenna-ground system to pick up as much as possible of the RF power passing through the air in its vicinity . In general, a higher antenna will pick up more power from the passing RF waves than will a lower one. (2) Convert the intelligence carrying AM sideband RF power into audio electrical power. (3) Convert the electrical audio power into sound power and get that power into the ear.

There are power losses at each of the three steps and our job is to minimize them in order to get as much of the sideband RF power passing through the vicinity of the antenna (capture area) changed into audio power for our ears. We want all of the "available power" at the antenna-ground system to be absorbed into the crystal radio set then passed on through it to our headphones as sound. However, some of it will be unavoidably lost in the RF tuned circuit. If the input impedance of the crystal radio set is not correctly matched to the impedance of the antenna, some of the RF power hitting the input to the crystal radio set will be reflected back to the antenna-ground system and be lost.

More information about measuring the Is you can find on the website of Ben Tongue , in his articles number 4 and 16.

I measured the Is value of several diodes, and calculated the diode resistance RD.
Also some European germanium types are measured.
Several diodes are measured of the type OA95 and AA119

Diode	VD (Volt) at 1 μA	Is (nA)	RD (kΩ).
HSMS282K	0.1341	7.9	3428
HSMS282K 2 parallel	0.118	14.5	1867
HSMS286K (1 diode)	0.1116	18.3	1479
5082-2800	0.1871	1.14	23756
5082-2835	0.1464	5.04	5373
5082-2835 2 parallel	0.1289	9.5	2850
BAT 82	0.136	7.3	3710
BAT 85	0.0686	90.8	298
OA81 (germanium)	0.0225	800	34

OA95 #1 (germanium) OA95 #2 OA95 #3 OA95 #4	0.0272 0.0221 0.0271 0.0304	600 821 604 502	45 33 45 54
AA116 (germanium)	0.0441	256	106
AA119 #1 (germanium) AA119 #2 AA119 #3	0.0320 0.0363 0.0428	461 370 272	59 73 100

The HSMS282K is the same as the HSMS2820, only the HSMS282K has 2 equal diodes in one package.
The Is value of the HSMS282K, the HSMS286K and the 5082-2835 is lower than the value in the datasheet, this has also been noticed by other people.

Ben Tongue wrote me, that the Is value of the 5082-2835 has been reduced over the years by the manufacturer.
Also temperature has big influence, I measured at 18 ºC, in datasheets the Is value is given at 25 flC.
A increase from 18 flC to 25 flC will increase the Is by 60 %.

power (1.0^2)/Ri to equal the output power [(1.0*2*sqrt2)^2]/Ro, the input RF resistance (Ri) must equal 1/8 Ro. That is, Ri=Ro/8. This illustrates the direct interaction between the RF input resistance and output audio resistive load. At high input power levels selectivity drops substantially if the output resistive audio load value is lowered.
The audio output resistance of the detector approaches 8 times the RF source resistance driving it. This fact is seldom recognized and it may be the cause of some of the problems encountered by those experimenting with doublers.

4) Half-wave voltage doubler operating at a low input signal power level: The detector, in this case, does not operate as a peak detector, and it has significant power loss. At low input signal power levels Ri approaches (0.026*n/Is)/2 ohms and becomes independent of the value of Ro.

The audio output resistance of the detector approaches twice the axis-crossing resistance of the diode.

5) Summary: At high input power levels, and with both input and output matched, power loss in both half wave and half wave voltage doubling detectors approaches zero dB. Sound volume should be the same with either detector. At low input power levels both detectors exhibit substantial power loss. I believe, but have not proven, that at low input power levels the doubler has a higher power loss than the straight half wave detector, and should deliver less volume.

6. Some misconceptions regarding Impedance matching and Crystal Radio Sets: To understand the importance of impedance matching, one must first accept the concept of power. A radio station accepts power from the mains and converts some of it to RF power which is radiated into space. This power leaves the transmitting antenna at the speed of light

1) Conventional half-wave detector operating at a high input signal power level: The detector, in this case, operates as a peak detector. Since it is a passive device, its output power will approximately equal its input power, under impedance-matched conditions. The output DC voltage will approach sqrt2 times the input RMS voltage, since the peak value of a sine wave is sqrt2 times its RMS value. For the input power, (1.0^2)/Ri, to equal the output power, [(1.0*sqrt2)^2]/Ro, the input RF resistance (Ri) must equal 1/2 Ro. That is, Ri=Ro/2. This illustrates the direct interaction between the RF input resistance and output audio resistive load. At high input power levels selectivity drops when the resistive audio output load value is lowered. The audio output resistance of the detector approaches 2 times the RF source resistance driving it. If the diode were an ideal diode, the word "approximately" should be eliminated, and "approaches" should be changed to "becomes" .

2) Conventional half-wave detector operating at a low input signal power level: The detector, in this case, does not operate as a peak detector, and exhibits significant power loss. At low input signal power levels Ri approaches 0.026*n/Is ohms (diode axis-crossing resistance) and becomes independent of the value of Ro.

The audio output resistance of the detector approaches the same value as the axis-crossing resistance (see above).

3) Half-wave voltage doubling detector operating at a high input signal power level: The detector, in this case, operates as a peak detector. Since it is a passive device, its output power will approximately equal its input power, under impedance-matched conditions. The output DC voltage will approach 2.0*sqrt2 times the input RMS voltage, since the peak of a 1.0 volt RMS sine wave is sqrt2 times its RMS value. For the input

Diode forward voltage drop:
David Knight

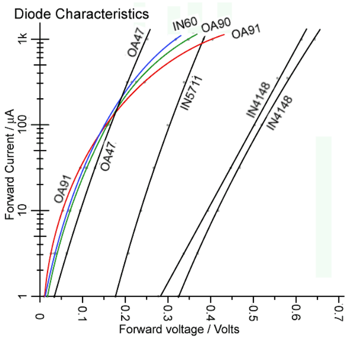
http://www.g3ynh.info/circuits/Diode_det.pdf

All of the diode detectors described above have at least one diode forward voltage-drop in the path to the measuring device. The diode forward voltage drop, Vf, varies approximately in proportion to the logarithm of the current, If, passing through the diode, and also depends on the temperature of the diode junction. For many types of diode, this behaviour is described to a good approximation by a modified form of the Ebers-Moll equation[12]:

$$V_f = m \, (kT/q) \ln[(I_f / I_s) + 1]$$

where k is Boltzmann's constant (1.380662 10 Joules/Kelvin) , q is the charge of an electron (1.6021892 10 Coulomb), and T is the temperature in Kelvin, i.e., the absolute temperature (flC+273.16). The factor kT/q is sometimes given the symbol VT, and has a value of 25.3mV at 20flC. "m" is a dimensionless correction factor between 1 and 2. "ln" is the natural logarithm (Loge). Is is the junction reverse saturation leakage current.

For small-signal rectifier voltmeters, diodes should be chosen for low forward voltage drop and low junction capacitance. Most sources of information now maintain that the best diodes in this respect are silicon Schottky-barrier (i.e., silicon-metal junction) diodes, such as the 1N5711 (Agilent 5082-2800) [18]; and so to investigate this matter, the forward voltage drop versus current characteristics of a variety of small signal diodes were measured. The results are shown in the graph below:



The 1N4148 is a silicon P-N junction diode. The 1N5711 is a silicon Schottky diode. The OA47 is an archaic germanium gold-bonded diode, and the rest are germanium point-contact diodes. Two 1N4148 diodes from different manufacturers were measured merely to illustrate the point that silicon P-N diodes have the highest forward voltage drop and are therefore a poor choice for low voltage detectors. The 1N5711 curve is the average of results from four diodes, all from the same batch, which had practically identical characteristics. The OA47 curve is the average for four diodes from two manufacturers, all having similar characteristics. The OA90, OA91, and 1N60 curves are from single examples, and are therefore not necessarily representative of the type. All data were recorded at an ambient temperature of 21flC.

signals are received, since both input and output ports become impedance matched.

Here is an interesting conceptual view of a high signal level diode detector circuit: Assume that it is driven with a sufficiently high level sine wave voltage so it operates in its peak detection mode, and is loaded with a parallel RC of a sufficiently long time const ant. This detector may be thought of as a low loss impedance transformer with a two-to-one impedance step up from input to output, BUT having an AC input and a DC output, instead of the usual AC input and output. The DC output power will approximately equal the AC input power and the DC output voltage will be about sqrt 2 times the RMS AC input voltage.

5A. A comparison of conventional half-wave and half-wave voltage-doubling detectors: Here is some info that may be of interest re conventional half-wave detectors vs. voltage doubling half-wave detectors when each is terminated with an output load of Ro. For illustration purposes we will assume the input voltage to the detector to be 1.0 volt RMS. The RF input resistance of the detector will be designated as Ri. All diodes have the same Is and n. It is assumed that good diodes such as a 5082-2835 Schottky, ITT FO-215 germanium or other are used. The info relates to the RF input resistance of detectors (it has a large effect upon selectivity) and their output audio resistance. See Point 4 in this Article for info on diode Is and n.

A high input power level is defined as one that is high compared to that at the LSLCP of the detector. A low input power level is defined as on that is low compared to that at the LSLCP of the detector. See "Quick Summary" in Article #15 for info on LSLCP.

resistance. Further, since the detector is now a peak detector, the DC output voltage is the "square root of 2" times larger than of the applied input RF RMS voltage. (It's equal to the peak value of that voltage). These existence of these relationships is necessary so that in an ideal peak detector, the output power will equal the input power (No free lunch). Summary: Output DC voltage equals sqrt2 times input RMS voltage. Since the output power must equal the input power, and power equals voltage squared divided by resistance, the output load resistance must equal two times the source resistance, assuming impedance matched conditions prevail. If we were to adjust the input source resistance to, say 495k ohms (reduce it by sqrt2) and the output load resistance to 990k ohms (increase it by sqrt2) by changing the input and output impedance transformation ratios, the insertion loss would become even lower than before the change and the input and output impedance matches would be very much improved (remember we are now dealing with high signal levels).

A good compromise impedance match, from one point of view, occurs if one sets the RF source resistance to 0.794*Rd and the audio load resistance to 1.26*Rd. With this setup, theoretically, the impedance match at both input and output remains very good over the range of signals from barely readable to strong enough to produce close to peak detection. A measure of impedance match is "Voltage Reflection Coefficient", and in this case it is always better than 18 dB (VSWR better than 1.3). Excess insertion loss is less than 1/3 dB and selectivity is largely independent of signal level. Information presented in Article #28 shows that, if the diode load resistance is made equal to Rd and the RF source resistance is made equal to Rd/2, the weak signal output of the detector will be about 2 dB greater than if both ports are impedance-matched! There is little benefit when strong

The data indicate that the IN4148, the 1N5711, and the OA47, all obey a logarithmic V/I relationship rather well, whereas the germanium point-contact characteristics show considerable curvature due to high internal resistance. With regard to the forward drop however, the germanium diodes are all superior to the IN5711 in the 1-100 A range, and the preference for the latter may merely reflect the fact that many semiconductor manufacturers can no longer fabricate germanium. Silicon Schottky diodes, such as the 1N5711 and 1N6263, being essentially UHF devices, have better high-frequency performance than germanium diodes, but germanium diodes work well at VHF and are therefore perfectly adequate for HF applications. Among the germanium diodes, there is little difference between the gold-bonded and standard varieties in the 1 to 100 A range, but the OA47 is the best choice for currents up to 1mA. We should observe however, that detector diodes only conduct on the peaks of the applied waveform, and so the instantaneous current is much higher than the average current, the difference being about an order of magnitude. Therefore, in selecting diodes for average currents in the region of 1 - 100 A, we should consider the steady-state voltage drop in the region 10 A - 1mA; in which case the germanium gold-bonded diode offers the lowest forward-drop without contest. Note however, that one of the consequences of the Ebers-Moll equation is that low forward voltage-drop is associated with high reverse leakage current. If reverse leakage is an issue, then silicon Schottky diodes are to be preferred. To put this issue in perspective however, the reverse leakage current of an OA47 diode was measured as follows:

Vr /Volts	1.0	2.0	3.0	5.0	10.0	15.0	20.0
Ir /uA	1.1	1.3	1.4	2.1	4.1	6.3	8.8

The leakage current is approximately linear in the 5-20V range and may be modelled by assuming a resistor of about 2M in parallel with the diode. Such a defect has little effect on the operation of a detector loaded with a 10-100K resistor. Some final points in favour of the silicon Schottky diodes however, are that germanium diodes show a wider spread of characteristics, and that the OA47 is obsolescent. Hence the Schottky diodes are definitely preferable in applications requiring diode matching, precise calibration, or availability through normal commercial channels. The 1N5711 in particular also, has a very high reverse breakdown voltage for a device of its class, its Vr max of 70V making it suitable for half-wave detectors of up to 24.7V DC ouput. An OA47 half-wave detector has a maximum DC output of 10.6V if Vr max is not to be exceeded.

Detector diode data: Source: refs. [18], [19]

having 3 dB less potential weak signal output than one of the diodes by itself.

5. Explanation of why, in a diode detector, and by how much, the RF input resistance and audio output resistances change as a function of input signal power. Consider first, a diode detector that is well impedance-matched both at its input and its output when driven by a very low power RF input signal. There will then exist an appreciable power loss in the detector. (The audio output power will be appreciably less than the input power.). The input and output impedances of the detector will approximately equal each other and approach: Rd = 0.026*n/Is. See part 3 above for a definition of terms. For this illustration, let the diode have an Is of 38 nA and an n of 1.02. Rd will be 700k Ohms. The well impedance-matched condition will hold if the input power is raised from a low value, but only up to a point. After that , the match will start to deteriorate. At an input power about 15 dB above that of the square-law-linear crossover point, the match will have deteriorated to a VSWR of about 1.5:1 (VSWR = Voltage Standing Wave Ratio.). A further increase of input signal power will result in a further increase of VSWR. This means that the input and output resistances of the detector have changed from their previously matched values.

The input resistance of the diode detector decreased from the value obtained in the well impedance matched low power level situation. The output resistance increased. The reason for this change is that a new law now governs input and output resistance when a diode detector is operated at a high enough power level to result in a low detector insertion power loss. It now operates as a peak detector. The rule here is that the CW RF input resistance of a diode peak detector approaches ½ the value of its output load resistance. Also, the audio output resistance approaches 2 times the value of the input AC source crystal radio set operation, but varies with diode current in silicon pn junction and germanium point contact diodes. A way of thinking about n is to consider it as a factor that effectively reduces the applied signal voltage to a diode detector compared to the case of using an ideal diode having an n of 1.0. Less applied signal, of course, results in less detected output.

Here are a few bits of information relative to diodes:
Typically, if a diode is biased at 0.0282*n volts in the forward direction, it will pass a current of 2 times its Is. If it is biased at 0.0182*n volts in the reverse direction, it will pass a current of 0.5 times its Is. If a diode is biased at 0.0616*n volts in the forward direction, it will pass a current of 10 times its Is. If it is biased at -0.0592*n volts, it will pass a current of -0.9 times its Is. These values are predicted from the classic Shockley equation. In the real world, reverse current can depart substantially from values predicted by the equation because of effects not modeled (the reverse current becomes higher). Gold bonded germanium diodes usually depart somewhat from the predicted values when operated in the forward direction. The effect appears as an increase of Is when measurements are made at currents above about 6 times the low-current Is.

Values of Is and n determine the location of the apparent 'knee" on a linear graph of the diode forward current vs. forward voltage. See Article #7. An easy way to estimate the approximate value of Is can be found in Article #4, section 2. A method of measuring Is and n is given in Article #16.

If one connects two identical diodes in parallel, the combo will behave as a single diode having twice the Is, and the same n as one of them. If one connects two identical diodes in series, the combo will behave as a single diode having twice the n and the same Is as one of them. This connection results in a diode

Detector diode data: Source: refs. [18], [19]

Type	Description	$V_{r\,max}$ / V	$I_{f\,max}$ / mA	$I_{f\,av}$ / mA	Typ V_f @ I_f / V	/ mA	Typ I_r @ V_r / µA	/ V
1N5711	Si Schottky (2.0pF)	70	-	-	0.41*	1	0.2	50
1N5712	Si Schottky (1.2pF)	20	-	-	0.55*	1	0.15	16
AA119	Ge point contact	45	100	35	2.6	30	170	45
AAY30	Ge Au-bonded. High speed	30	400	-	0.88	150	8.0	30
AAY32	Ge Au-bonded. High speed	30	150	-	0.60*	30	11	30
AAY33	Ge Au-bonded. High speed	12	240	-	0.5*	30	15	12
AAZ15	Ge Au-bonded. High voltage	100	250	-	0.8	250	16	100
AAZ17	Ge Au-bonded. Gen.purpose	75	250	-	0.8	250	16	75
BAT81	Si Schottky (1.6pF)	40	30	-	0.33	0.1	0.2	30
BAT82		50			0.41	1		
BAT83		60			1.0	15		
OA47	Ge Au-bonded. Gen.purpose	30	150	-	0.54	30	10	30
OA90	Ge point contact "diode"?**	30	45	10	2.0	30	300	30
OA91	Ge point contact	115	150	50	2.1	30	75	100
OA95	Ge point contact	115	150	50	1.85	30	80	100

* max.
** a very non-linear resistor pretending to be a diode - best avoided!

6-8a. Thermionic diodes:
Given that semiconductor diodes are imperfect, some readers may wonder if thermionic diodes (valves) are capable of better performance. The answer however (avoiding expletives) is an an unequivocal 'no!'. The forward voltage of an indirectly heated valve diode in the low-current (space-charge limited) regime (according to Terman [20]) is given by the expression:

$$V_f = [\sqrt[3]{(I_f^2 / k)}] - V_c$$

where k is a constant determined by the geometry of the diode and Vc is called the contact potential, by analogy with the potential developed by a thermocouple. The diode contact potential arises because some of the electrons ejected from the cathode arrive at the anode even when there is no bias, and so the anode becomes negatively charged. This means that the diode must be reverse biased in order to prevent it from conducting, and the amount of bias required varies with the heater temperature and the age of the valve. In precision measurement terms, the contact potential is enormous. A sample of eight double-diode valves of type EB91 (6AL5) showed contact potentials ranging from 0.48 to 0.96V (with a mean of 0.73V) when measured using a voltmeter with 10M input resistance and a stabilised 6.30V DC heater supply (sensible measurement was impossible using a filament transformer connected to the domestic mains supply). A 100 A moving coil meter with an internal resistance of 980 was connected across a diode having an open-circuit contact potential of 0.74V, and registered a zero-bias current of 70 A, i.e., the diode gave a DC output of 4.8 W due to the thermal current. When the meter was padded to 10K , to simulate the diode loading in a realistic detector circuit, the zero-bias current was 22 A, i.e., 22% of full-scale deflection. We may safely conclude that thermionic diodes have no merit whatsoever in small-signal measuring applications, even if they do glow attractively in the dark.

(Ed note: thermionic diodes were the original diode detectors and can STILL make an interesting radio.)

increase in reverse current at a specified voltage. These diodes are called Zener diodes.

Diode Saturation Current is a very important SPICE parameter that, along with the diode Ideality Factor, n determines the actual diode current when it is forward biased by at particular DC Voltage. Id=Is*(e^(Vd/(0.026*n)-1) at room temperature. This expression ignores the effect of the parasitic series resistance of the diode because it has little effect on the operation of crystal radio sets at the low currents usually encountered. Here Id is the diode current, e is the base of the natural logarithms (2.7183...), ^ means raise the preceding symbol to the power of the expression that follows (Sometimes e^ is written 'exp'), * means multiply the preceding and following symbols, VD is the voltage across the diode and n equals the "Ideality factor" of the diode. At low signal levels, most detector diodes have an n of between 1.05 and 1.2). The lower the value of n, the higher will be the weak signal sensitivity. One can see that Is is a scaling factor for the actual curve generated by the factor (e^(VD/(0.026*n)-1).

Diode ideality factor (n): The value of n affects the low signal level sensitivity of a diode detector and its RF and audio resistance values. n can vary between 1.0 and 2.0. The higher the value of n, the worse the low signal level detector sensitivity. The low signal level RF and audio resistances of a diode detector vary directly with the value of n. Schottky diodes usually have a value of n between 1.03 and 1.10. Good germanium diodes have an n of about 1.07 to 1.14 when detecting weak signals. Silicon p-n junction diodes such as the 1N914 have values of n of about 1.8 at low currents and therefore have a lower potential sensitivity as diode detectors than Schottky and germanium point contact diodes. The value of n in Schottky diodes seems to be approximately constant over the full range of currents and voltages encountered in source, so we can say we have a 'no loss' situation. Now, assume that a transformer or other two-port device is inserted between the Vs,Rs source and Ro, and that an output voltage Vo is developed across Ro. The output power is (Vo^2)/Ro. The 'insertion power loss' can now be calculated. It is: 10*log (output power)/(maximum available input power) dB. After substituting terms, the equation becomes: Insertion power loss =10*log [(Vo/Vs)^2)*(4*Rs/Ro)] dB.

If the input voltage is referred to by its peak value (Vsp) as it is in a SPICE simulation, instead of by its RMS value, the equation changes. The RMS voltage of a sine wave is equal to the peak value of that wave divided by the "square root of 2". Since the power equation squares the voltage, the equation for the 'available input power' changes to Pa = (Vsp^2)/(8Rs).

4. Diode Saturation Current and Ideality Factor: Saturation current is abbreviated as Is in all of these articles. Assume that one connects a DC voltage source to a diode with the polarity of the voltage source such as to bias the diode in the back direction. Increase the voltage from zero. If the diode obeys the classic Shockley ideal equation exactly, the current will start increasing, but the increase will flatten out to a value called the saturation current as the voltage is further increased. That is, as the voltage is increased, the current will asymptotically approach the saturation current for that diode. A real world diode has several mechanisms that cause the current to actually keep increasing somewhat and not flatten out as the back direction voltage is further increased. Diode manufacturers characterize this as reverse breakdown and specify that the back current will be less than a specified value, say 10 uA at a specified voltage, say 30 V, called the reverse breakdown voltage. BTW there are other causes of excessive reverse current that are collectively referred to as reverse bias excess leakage current. Some diodes have a sharp, controlled

A Procedure for Measuring the Saturation Current and Ideality Factor of a Diode, along with Measurements on various diodes
http://www.bentongue.com/xtalset/16MeaDio/16MeaDio.html
by Ben H. Tongue

Quick Summary: A schematic and operational instructions are given for a device for use in measuring Saturation Current and Ideality Factor of a diode. Measurements of various detector diodes are included.

The Saturation Current and Ideality Coefficient of a diode can be determined by measuring an applied junction voltage along with the associated current flow at two different voltages. These two data pairs are then substituted into the Shockley diode equation to create two simultaneous equations in Is and n, and then solved for Is and n. Since the equations include exponential functions, they can not be solved by ordinary algebra. Numerical methods must be used.

The Shockley diode equation at 25 degrees C. is: Id = Is*(exp(Vd/(0.0256789*n))-1) Amps. Id = Diode Current (amps), Is=Saturation Current (amps), Vd = Diode Voltage, n = Ideality Coefficient. The series resistance Rs of the diode is ignored because the measurement currents are so low that the voltage drop across Rs is negligible. Measurements have shown that Is and n of point contact germanium diodes can vary with current, but are relatively constant, down to very low currents, when the current is under six times Is. Silicon p-n junction diodes exhibit values of Is and n that vary with current. The values for Is and n of Schottky diodes are quite constant over the range of currents used in ordinary crystal radio set reception.

A convenient set of measuring currents is about 6*Is and 3*Is. Substituting Id = 6*Is, then Id = 3*Is into the Shockley and solving for Vd yields: For Id = 6*Is, Vd = 0.05000*n volts. For Id = 3*Is, Vd = 0.03561*n volts. The value of n will probably be between 1.0 and 1.2 for the type of diodes used in crystal radio sets, so use 1.1 in determining the applied voltage to use. Suggested voltages to use are about 0.055 and 0.039 volts, although other values may be used.

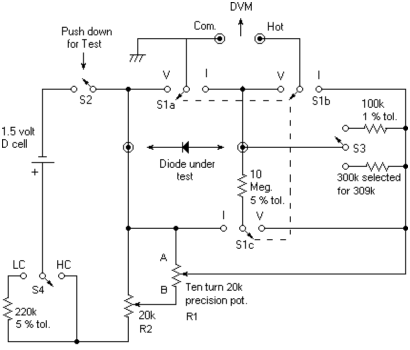


Fig. 1 - Schematic of Device for Measuring Diode Is and n.
Schematic

S1 is a triple pole double throw switch, S2 is a push button momentary-contact SPST switch. DVM is a digital voltmeter with 10 Meg input resistance having a 200 mV range setting. S3 is a range switch that enables greater precision when using a conventional 3 1/2 digit DVM. It is also used when

gain of, say 80 times, the output power is 2.5 X 80=200 mW. 2.5 mW expressed in dBm is +4 about dBm. A power gain of 80 times is about +19 dB. The output power is 4+19=+23 dBm.

3. Maximum Available Power: If one has a voltage source Vs with an inaccessible internal resistance Rs, the load resistance to which the most power (Pa) can be delivered is equal to Rs. Pa is called the 'maximum available power' from the source Vs, Rs. Any load resistance other than one equal to the source resistance, Rs, will absorb less power. This applies whether the voltage is DC or AC (RMS). The formula for power absorbed in a resistance is "voltage-squared divided by resistance". In the impedance matched condition, because of the 2 to 1 voltage division between the source resistance and load resistance, one-half of the internal voltage Vs will be lost across the internal source resistance. The other half will appear across the load resistance. The actual power available to the load will be, as indicated in the preceding relation: Pa = [(Vs/2)^2]/Rs = (Vs^2)/(4*Rs). Again, in the impedance matched condition, the total power delivered to the series combination of source and load resistance is divided up into two halves. One half is unavoidably lost in the internal source resistance. The other half is delivered as "useful output power" to the load resistance.

The 'maximum available power' approach is useful when measuring the insertion power-loss of two-port devices such as transformers, amplifiers and crystal radio sets, which may not exhibit an input or output impedance that is matched to the power source. The input impedance may be, in fact a combination of resistive and reactive components. If the Vs,Rs source is connected to a resistive load (Ro) of value equal to Rs ohms, it will receive and dissipate a power of Pa Watts. This is the maximum available power from the Vs, Rs frequencies. This is the main reason why diodes having large values for Rs and CJ perform poorly at high frequencies.

2. An explanation of the meaning and use of dB and dBm: In the acronym dBm, "d" means one-tenth. "B" refers to the Bel and is named after Alexander Graham Bell. The Bel is used to express the ratio of two powers, say (Output Power)/(Input Power). Let's call this power ratio "(pr)". Mathematically, a power ratio, expressed in Bels, is equal to the logarithm of the ratio of the two powers. B=log (pr). If the two powers are equal, the power ratio expressed in Bels is 0 B. This is because the log of one is zero. Another illustration: Assume that the power ratio is twenty. (Pr)=20. The logarithm of 20 is about 1.3. This power ratio in Bels is 1.3 B. One decibel is equal to 0.1 Bel. That is, 10 dB=1 B. If we express the two power ratios mentioned above (1 and 20) in dB, we get 0 dB and about 13 dB.

So far, we have seen that the decibel is used to express the ratio of two powers, it is not a measure of a power level itself. A convenient way to express an actual power level using dB is to use a standard implied reference power for one of the powers. dBW does this. It expresses the ratio of a power to the reference power (One Watt in this case). dBm uses a reference power of one milliwatt. A power level of, say 100 milliwatts, can be said to be a power level of +20 dBm (twenty dB above one milliwatt). Why? (100 milliwatts)/(1 milliwatt)=100. The logarithm of 100 is 2. 10 times 2 equals 20.

The convenient thing about using dB comes from a property of logarithms: The logarithm of the product of two numbers is equal to the sum of the logarithm of each number, taken separately. An illustration: If one has a power source of, say 2.5 mW and amplifies it through an amplifier having a power

measuring diodes having a high Is. R2 is used for coarse setting of the diode voltage. R1 is a ten turn precision 20k pot such as part # 594-53611203 from Mouser. It is used for fine setting of the diode voltage.

Procedure for Measuring Is and n:

1. Set S3 for 300k for diodes expected to have a low to medium Is. Set S3 to 100k if the diode is expected to have a high Is. S4 to HC and R1 to 1 about turn from point B.
2. Take Data Set #1: Set S1 to V. Push S2 and adjust R2 to obtain a reading of about 0.055 volts. Use R1 to set the voltage to the voltage desired (0.055 volts is suggested). Call this voltage V1. Set S2 to I, read the DVM and call that voltage V2.
3. Take Data set #2: Set S1 to V. Push S2 and adjust R2 to obtain a reading of about 0.039 volts. Use R1 to set the voltage to the voltage desired (0.039 is suggested). Call this voltage V3. Set S2 to I, read the DVM and call that voltage V4.
4. The diode voltage (Vd1) from Data Set #1 is V1. The diode current from Data Set #1 (Id1) is (V2/300,000)-(V1/10,000,000) or (V2/100,000)-(V1/10,000,000) Amps, depending on the setting of S3. The diode voltage (Vd2) from Data Set #2 is V3. The diode current (Id2) is (V4/300,000)-(V3/10,000,000) or (V4/100,000)-(V3/10,000,000) Amps, depending on the setting of S3.
5. The two data sets Vd1, Id1 and Vd2, Id2 must now be entered into two Shockley diode equations (shown above) in order to make two simultaneous equations in Is and n. Solving them will yield values for Is and n, measured at an average current of about 4.25 times Is.

A numerical equation solver can be used to solve the two simultaneous equations for Is and n. One is available in

MathCad. If you have MathCad 5.0 or higher, go to http://www.agilent.com/. Click your way through Communications, Communications Designer Solutions, RF and Microwave, Schottky Diodes, Library, MathCad worksheets and download the file: sch_char.mcd. Execute it in MathCad, then enter your Current and Voltage values: Id1, Vd1 and Id2, Vd2 as I2, V2, I1 and V1. Pull down 'Math' and click 'Calculate Worksheet'. The program calculates Is and n. Since most crystal set operation occurs at currents so low that there is negligible voltage drop across the diodes' parasitic series resistance, there is no need to enter any new numbers for I3, 4, 5 and V3, 4, 5 on the worksheet. The program sch_char.mcd does not work in versions of MathCad earlier than 6. If you have an earlier version of MathCad, and it has a non-linear equation solver, actual entry of the Data Set will have to take place without the convenience of the sch_char program. Those who do not have MathCad but do have Microsoft Windows Word can get an unformatted view of the default data and text provided in the MathCad program by clicking here.

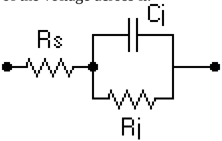
There is currently available on the Web, a program from Polymath Software at: http://www.polymath-software.com/. This program has many capabilities, and among them is a nonlinear equation solving capability. A free demo copy of the latest program is available for download, but is limited to 20 uses. After that, for more usage, you have to buy it.

Some programmable pocket calculators include a nonlinear equation solver. One calculator that has one is the HP 32S Scientific Calculator. A program to solve for n and Is takes only 28 steps of program memory and is here.

Mike Tuggle posted on 'The Crystal Set Radio Club' the following simple procedure for determining Is and n by using a

PRACTICAL CONSIDERATIONS, helpful definitions of terms and useful explanations of some concepts
Ben H Tongue
http://www.klimaco.com/HAMRADIOPAGES/xtal_how_to.htm

1. An explanation as to why some diodes that work well in the Broadcast Band cause low sensitivity and selectivity when used at Short Waves: The parasitic (approximately fixed) series resistance Rs of a diode is in series with the parallel active elements. The nonlinear active elements are the junction resistance Rj, which is a function of current through the diode, and the junction capacitance Cj, which is a function of the voltage across it.



The nonlinear junction resistance effect is what we use to get detection. The nonlinear capacitance effect is used when the diode is designed to be a voltage variable capacitor (a varactor diode).

The parasitic series resistance of some 1N34 diodes can be pretty high, and in series with the junction capacitance, make that capacitance have a rather low Q at high frequencies. This capacitance is, in a crystal radio set, effectively in parallel with the tank. The tank usually has a small value tuning capacitor itself, so the overall tank circuit Q is reduced at high resistive loading occurs from the real but usually neglected current passed in the back direction during the RF reverse peak voltage excursions. This causes power loss and is a cause of loss of volume. For a given Is, good germanium diodes are usually superior to Shottkys in this respect.

Best regards,

Ben
eug4not

spreadsheet. 'In lieu of an equation solver package, the Schottky parameters can be solved for by simple trial-and-error. This is easily done with an ordinary spreadsheet, like Excel or Lotus. For the two measurement points, (Id1, Vd1) and (Id2, Vd2), set up the spreadsheet to calculate: Id2[exp(Vd1/0.0257n) - 1] and, Id1[exp(Vd2/0.0257n) - 1]. Then experimentally plug in different trial values of n, until the two expressions become equal. This gives the correct value of n. Now, plug this value of n into: Is = Id1 / [exp(Vd1/0.0257n) - 1] or, Is = Id2 / [exp(Vd2/0.0257n) - 1] to get the correct value of Is.' An Excel spreadsheet constructed as Mike suggested is here. An example from data taken on an Agilent HBAT-5400 is entered, for reference, on line 2. Line 3 may be used for calculations using data from other diodes. Column H automatically calculates a value for Is each time n is changed. All one has to do is enter the values as described above in columns A through E and hit enter.

Caution: If one uses a DVM to measure the forward voltage of a diode having a high saturation current, a problem may occur. If the internal resistance of the DC source supplying the current is too high, a version of the sampling voltage waveform used in the DVM may appear at its terminals and be rectified by the diode, thus causing a false reading. One can easily check for this condition by reducing the DC source voltage to zero, thus leaving only the internal resistance of the source in parallel with the diode, connected across the terminals of the DVM. If the DVM reads more than a tenth of a millivolt or so, the problem may be said to exist. It can usually be corrected by bypassing the diode with a ceramic capacitor of between 1 and 5 nF, preferably, an NPO type. I use a 0.047 uF NPO multi-layer ceramic cap from Mouser Electronics. Connect the capacitor across the diode with very short leads, or this fix may not work.

Tips

* If the Is of the diode under test is too high, 0.055 volts will not be attainable for V1 in step 1. The solution is to set switch S3 to 100k. The calculations for diode current then become: Id1=(V2/100,000)-(V1/10,000,000) Amps and (Id2=V4/100,000)-(V3/10,000,000) Amps.

* If the voltage readings seem to unstable, try placing the measuring setup on a ground plane and connect the common lead of the DVM to it. A sheet of household aluminum can be used for the ground plane. Use shielded cable from the lead from the DVM to the test setup.

* The voltage readings are very sensitive to diode temperature. You can see this easily by grasping the diode body with thumb and forefinger and noting the change in the voltage reading when measuring V1 or V3. Don't take data until the readings stabilize. Saturation current is a strong function of junction temperature. For germanium and the usual (n-doped) Schottky diodes, a temperature increase of 10fl Celsius results in a saturation current increase of about two times. A simple rule is: For each 1fl C. increase in temperature, Is increases by 7.2%. The figures are different for zero-bias-type Schottkys. Here, a 14 degree C. (25 degree F.) change in temperature will result in approximately a two times change in Is.

* Shield glass enclosed diodes from ambient light by placing a cardboard box over the unit. Many diodes have a photo-diode response and will give an output voltage when exposed to light even if no current is applied.

Note: A simplified method of determining the Saturation Current of a diode, if the Ideality Factor is estimated in advance is shown in Section #2 of Article #4.

resonance increases as the coupling between the detector and ATU is reduced - and therefore a diode with a larger rc value is a better match. If this is the case, then for really good radios with very high Q detector tank circuits, two different diodes may give the loudest output - one when coupling is very tight, and a different one when coupling is very loose. (Any comments??)

For my radio used here, the 1N141 diode (thank you Bruce Kizerian!) seems to give about the loudest output for both tight and loose coupling at the low end of the band. But the FO-215's output is very close and has a bit better selectivity. The schottky diodes (1SS98 & 5082-2835) are seen to be useful even for my modest radio if there is plenty of signal that can be traded-off for greatly improved selectivity.

Well, that's my attempt at measuring my crystal radio's audio output. Hope it's of interest. If my measurements or methods are way off base, please let me know.

It sure is a blessing to be retired and to have the luxury of time to play.

73's, Dan

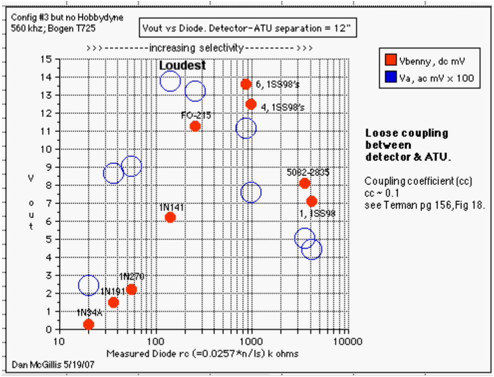
May 23, 2007
Hi Dan,

You have developed some really insightful information.

There is one thing, I believe, that is being overlooked re the low Is (high Rx) Schottky diodes when one is going for maximum volume. They have rather low reverse breakdown voltages so, when a large RF is being detected, a lot of extra

2.) Loose coupling between the detector and ATU (12" separation).

The last plot again shows how the ac "audio" voltage Va and the dc benny voltage Vb vary as different diodes are used in the radio.



The benny voltage (RED DOTS) seems to peak in the vicinity of the benny resistance as before. However, the peak ac "audio" voltage Va range (BLUE CIRCLES), may be shifting toward higher diode rc values than for the tight coupling case.

My assumption is that the detector tanks' parallel resistance at

Summary of measurements on some diodes:

The following charts show typical values for Is and n for diodes that might be used in crystal radio sets. One can see, for any particular diode, that Is and n do not vary by much over a moderate current range. Therefore, they may be considered to be dynamically constant when receiving a signal. Each value of n and Is is calculated from two voltage/current pairs as described above. The diode current (Id) given for each of the n, Is pairs is the geometric mean of the two currents used in the measurement. A Fluke model '89 IV' 4-1/2 digit DVM was used to enable measurements down to as low as 15 nA on some diodes. Noise problems cause some measurement error at low currents. That is the reason for the fluctuations in some of the readings. Values of n very close to 1.0 or below are obvious measurement errors. Those low values for n should have come out somewhat higher and the associated values of Is, also higher.

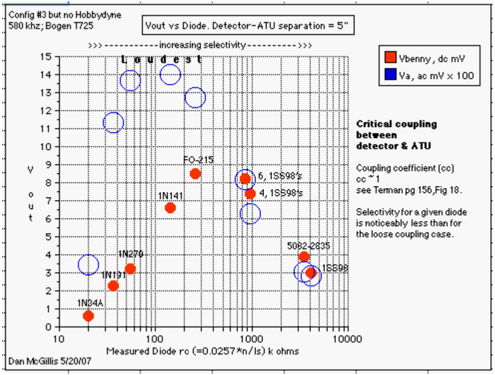
Note that the germanium diodes show an unexpected tendency to increased values for Is and n at the higher currents. The 1N4148 silicon p-n junction shows the expected increase of Is and n at lower currents. The Schottky diodes seem to have pretty constant values of Is and n across the current ranges measured. Experiments described in Article #27 indicate that the measured values of Is and n for silicon Schottky diodes tested here, when used as detectors, remain at the measured values at rectified currents so low that a voice signal is barely readable. This is not necessarily true for all germanium diodes.

Table 1 - Measured n and Is values for various diodes, over a range of currents (Id), in nA.

1N4148 silicon p-n junction diode			Base-emitter junction of 1N404A Ge transistor			Blue Radio Shack 1N34A Ge diode having no nomenclature			Agilent HBAT-5400 Schottky, high Is version			Infineon BAT62-03W Schottky diode		
Id	**n**	**Is**	**Id**	**n**	**Is**	**Id**	**n**	**Is**	**Id**	**n**	**Is**	**Id**	**n**	**Is**
710k	1.73	1.23				710k	1.71	3500						
			570k	1.04	1800									
350k	1.75	1.45				350k	1.69	3200						
177k	1.87	2.19	179k	1.03	1670	177k	1.61	2550						
88k	1.82	2.26				88k	1.51	1980						
44k	1.80	1.98	56k	1.01	1540	44k	1.39	1470						
22k	1.88	3.00	18.9k	1.01	1580	22k	1.28	1100						
11k	1.89	3.10				11k	1.22	950						
5500	1.93	3.80	5700	1.04	1660	5500	1.14	800	8100	1.15	265			
2760	1.94	3.90	1790	0.98	1730	2760	1.10	750				2600*	1.06	248
1380	2.02	4.90				1380	1.05	680						
690	1.98	4.40	620	0.99	1740	690	1.20	830	990	1.15	248	970	1.04	240
343	2.06	5.30				343	1.01	670	360	1.15	265	341	1.04	236
170	2.18	6.70				170	1.08	720	160	1.15	255	133	1.04	236
									76	1.15	254	87	1.04	236
												59	1.01	228
									40	1.15	261	39	1.06	233

* This Infineon diode has an unusually high series resistance of 130 ohms. The voltage drop across this resistance is low enough in all the measurements to be ignored, except for the highest current one. There, a correction for the voltage drop was made.

Table 2 - Measured n and Is values for various diodes, over a range of currents (Id), in nA.



If Va is in fact related to audio power, then the plot indicates that for this radio, the audio power delivered peaks for diodes with critical resistances in the range of say 50k to 200k - a pretty broad range. The dc benny voltage Vb (RED DOTS) however, seems to peak in the vicinity of about 500k - the value of the benny resistance. As others have noted, the selectivity - as judged by listening - continuously improves as diodes with larger and larger rc values are used. By choosing different diodes, you can tradeoff loudness for selectivity. The plot shows the magic of impedance matching. (Thank you Ben Tongue for your articles!) All this is old hat to some folks but for a novice like me, it's neat to see.

The modulation on my talk show test signal seems fairly constant - as long as measurements are avoided during commercials or the occasional music. A spotting radio helps in avoiding high modulation surprises. The ac signal does still bounce around a bit. Taking 20 random voltage readings per data point and averaging helps smooth the results.

The radio used as a test bed is double tuned with 165/46 litz coils close-wound directly on Amidon 0.5”x4” ferrite 61 rods. The coupling between the detector’s tank circuit and the ATU tank is varied by changing the end-to-end separation between the ferrite rods.

1.) Tight (approximately critical) coupling between the detector and ATU.

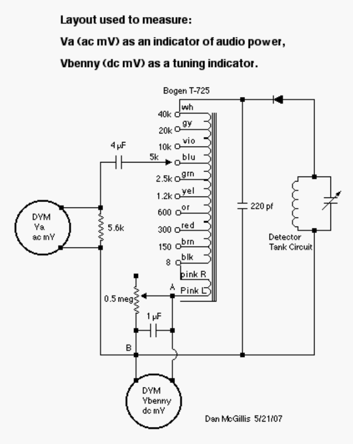
The first plot shows how the “audio” ac voltage Va (BLUE CIRCLES) varies as the detector diode is changed - with a small 5” separation between the detector and ATU rods.

Radio Shack Ge 1N34A diode marked 12101-3PT			**Agilent HBAT-5400 Schottky diode (low Is version)**			**Agilent HSMS-282M quad Schottky, all four diodes in parallel**			**Agilent HSMS-286L triple Schottky, all three diodes in parallel**			**One diode of Infineon BAT62-08S triple diode Schottky**		
Id	**n**	**Is**	**Id**	**n**	**Is**	**Id**	**n**	**Is**	**Id**	**n**	**Is**	**Id**	**n**	**Is**
47k	1.28	230												
17k	1.18	188												
9.5k	1.16	174												
			6.7k	1.03	102							4650	1.03	143
2.8k	1.13	160												
						1140	1.03	47	1750	1.04	76			
630	1.15	162	510	1.03	104	470	1.02	41				700	1.02	142
									360	1.04	76			
205	1.15	166				203	1.02	41				222	1.02	136
			151	1.03	103	108	0.98	40	117	1.02	72			
81	1.14	161										99	1.01	134
			59	1.01	102	59	0.99	39				47	1.02	138
37	1.13	160				36	1.00	39	53	1.03	73			
			26.4	1.02	100	23.1	0.98	39	24.5	1.02	72	23.6	1.01	135
			13	1.03	102	15.3	1.02	40						
						10.2	1.00	39	12.6	1.06	76			
						6.3	1.08	42						
						4.4	1.04	42						

A rare germanium diode that seems to be ideal for many crystal radio set designs is the FO 215, branded ITT. A search of the Internet has not turned up a manufacturer's datasheet. ITT is not in the germanium diode business anymore, but from the Internet search it appears that the original company was a German company named ITT Intermetall. Some of their semiconductor business became ITT Semiconductors. This was later sold, around 1997 to General Semiconductor Industries. That business was later sold to Vishay. One source indicated that General Instruments was also one of the intermediate owners. Averages of measurements on three samples of the FO 215 are: Is=109 nA and n=1.02. These measurements were made at an average current of about 250 nA. Interesting note: The average Is of the FO 215 diodes is about equal to the geometric mean of that of the Agilent 5082-2835 and a typical 1N34A. I obtained my FO 215 diodes from Mike Peebles at: http://www.peeblesoriginals.com/ .

Article #27 shows detector measurements of how diodes having different values of Is and n perform as weak signal detectors when impedance matched at both input and out put.

#16 Published: 03/28/01; Revised: 02/10/2004



The ac voltage (Va) across the 5.6k resistor load is assumed to be proportional to audio power (=V^2/R). The DVM used is supposed to have a 10 meg input on the “ac mV” scale. It’s specs say it has a 40 - 400 hz ac response, and an rms output calibrated to a sine wave. Since the audio voltage across the 5.6k is not a sine wave, and has frequency components outside the 40 - 400 hz response range, I can only hope that the DVM’s output will be a relative measure of audio power. The DVM is all I’ve got, so it will have to do.

Audio and Benny Voltage Measurement:
May 21, 2007

A lot of the fun playing with crystal radio for me involves "measuring stuff". The "benny voltage" - the dc voltage across the benny resistor in the matching transformer circuit - is a favorite parameter of mine. It's easy to measure with a DVM and it seems to vary with everything. And it's a very good tuning aid.

However, Dave Schmarder has gently and patiently reminded me on many occasions that it's the audio power delivered to the headphones that's important -- that I should try to measure relative audio power as well as benny voltage when evaluating radio changes. The following is my first attempt. Dave thought it might be of interest to the group.

With a DVM as the only measurement tool, and weak daytime talk stations (non sports) as signal sources, my measurements are low-tech. The circuit used to measure the ac voltage (Va) - that I hope is related to relative audio power - is shown in the attached layout.

Diode Modeling Update
Kevin Smith
http://www.lessmiths.com/~kjsmith/crystal/dmodel.shtml

The following explains my first attempt to compare a measured diode characteristics with calculated plots based on the Shockley Equation.

I have been taking serious look into diode modeling and measurement recently in an attempt to better understand these solid-state successors to the humble crystal and cat's whisker detector. There is much good information on diodes to be found in many excellent web sites of course, but there is nothing quite like actually making the measurements and working with them to get a good feel. This page reports some of my protocols and test setups which I have found to be useful.

Most probably the single best web article for the measurement of diode Is and n parameters is given by Ben Tongue in his Article 16. In this article he describes an interesting circuit for the measurement as well as protocols for making them. For my purposes I did not wish to make this a construction project and felt I might get along with good quality meters and my already-built "Diode Test Jig". Essentially Ben Tongue's method consists of making two precision measurements of Voltage and Current through a diode at small signal levels, essentially about 3 and 6 times Is (sufficiently low that the voltage drop across the series resistance Ro can be ignored). The measurements are then substituted into the Shockley equation Is*{exp[(qe/(n*k*T))*(V-I*Rs)]-1} and solved simultaneously for the two data pairs. Mike Tuggle has provided a nice excel spreadsheet to do the math. This spreadsheet, Cal_n_Is.xls forms the basis of my technique and

I am thus indebted to both Ben Tongue and Mike Tuggle for making this project feasible.

In addition to simply calculating the main parameters Is and n, I also wished to measure enough points to plot a characteristic curve, and to compare the measured curve with one calculated directly from the Shockley equation. My first spreadsheet then combined Tongue/Tuggle calculation protocol and a graphical view of the match between measured and theoretical. I did make a few methodology modifications which I thought/hoped would allow added accuracy:

1) Using the full Shockly equation without simplifying for an assumed 25dC room temperature. While the 25dC assumption is generally good and probably within the measurement error (or perhaps not), including the actual measured temperature eliminates doubts of inaccuracies due to this parameter. With the power of PC's and spreadsheets there is no reason simplify the equation.

2) Used Vd = 0.04 and 0.05 V (adequately close to Ben Tongue's recommended 0.039 and 0.055 V) whenever possible. Note for some diodes with significantly different forward voltage drops, Si diodes in partictular I had to use higher values for Vd.

3) Reporting: I decided that, as there is no unique solution to n and Is, each is dependant on the valuse Vd and Id, I feel that reporting both Vd and Id is necessary for repeatability. Naturally I also include the ambient temperature in the report as well, and

4) I included a calculation for Ro (or Rg if you are using Tongue's reference) as that is the goal of the exercise.
A screen shot shown below:

Effect of Diode Rd on Audio Power
http://theradioboard.com/rb/viewtopic.php?t=330
By Dan McGillis

Introduction:
Sep 13, 2011

Kevin - thanks for posting your work. Really good stuff.
Mike Tuggle & Mike Peebles gave excellent advice on choosing a diode.

Here's a plot of how the choice of diodes - and their associated Rd's - affected the relative loudness of one of my crystal sets.
http://theradioboard.com/rb/viewtopic.php?t=330

As you can see, there is a "best" choice of Rd for this particular set if maximum loudness is the goal. Using diodes with Rd's larger than that needed for maximum loudness gave the crystal set more selectivity (and less loudness).

I should note that you get this kind of curve of loudness vs diode Rd when detecting WEAK signals. For very strong signals, all my diodes gave about the same measured loudness. ie. The choice of diode (& it's Rd) doesn't matter too much for very strong signals.

As a side note, my measured Rd's for a given diode are a bit higher than yours and Mike's because I tend to measure at higher Vd's for convenience. And my measurements are no where near as accurate as yours.

Sure is fun stuff isn't it.

73, Dan

then: d2I / dV2 = d/dV [U-1] = -1 * U-2 * dU / dV, where

dU / dV = (dU / dV) + (dU / dI) * (dI / dV), and where

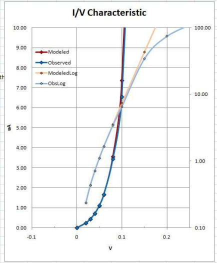
(dU / dV) = (-1 / Is) * exp [-q (V - I Rs) / nkT]

(dU / dI) = (Rs / Is) * exp [-q (V - I Rs) / nkT]

Sze's Gamma:

γ = (d2I / dV2) / (dI / dV)

FO-215						
	Is =		233	nA	2.33E-07	Amps
	n =		1.11	Id =	0.9130	k nA
	Id =	Is*{exp[(qe/(n*k*T))*(V-I*Rs)]-1}				
	Temp			Id		83.1
	28.4	dC	obs	modeled		dF
	Vd		uA	Amps	uA	
	-0.02			-1.17E-07	-0.1171	
	0		0.0000	0.00E+00	0.0000	Tuggle Meth
	0.02		0.2375	2.35E-07	0.2350	
	0.03		0.4370	4.30E-07	0.4301	3.36E-06
	0.04	Vd1 Id1	0.7149	7.07E-07	0.7066	3.365E-06
	0.05	Vd2 Id2	1.1110	1.10E-06	1.0983	calc ratio
	0.06		1.6610	1.65E-06	1.6532	1.000
	0.08		3.4310	3.55E-06	3.5531	
	0.10		6.5310	7.37E-06	7.3664	
	0.15		34.0900	4.31E-05	43.1395	
	0.20		74.3600	2.47E-04	247.3	
	0.30		180.29	8.06E-03	8062.6	
	0.40		1258.00	2.63E-01	2.63E+05	
	0.50		11310.00	8.55E+00	8.55E+06	
	0.60		18234.00	2.79E+02	2.79E+08	
	0.70			9.08E+03	9.08E+09	
	0.80			2.96E+05	2.96E+11	
	0.90			9.63E+06	9.63E+12	
	T dK	Vd av mV	Id av nA	n	Is nA	
	301.4	45	913	1.11	233	



To use the sheet one need only adjust the voltage Vd to that shown in the first column and record the current Id. If only Vd1/Vd2 and Id1/Id2 are measured then the next step is to adjust the value for "n" until the match between the two equations is exact. I even provided a simple ratio calculation to easily test for a match.
While the math is good and results excellent, I quickly found a couple deficiencies in my method,
1) measurements require high precision and its nearly impossible to land EXACTLY at the voltage required. I needed only to get quite close, let the meters stabilize for 2 - 5 minutes, and then record BOTH Vd and Id.
2) the work is rather tedious and, for a large number of diodes it pays dividends to measure only the needed Vd and Id and let the plot aside. My second spreadsheet thus dispenses with the plot. It is with my second spreadsheet that all my data tables and results are posted.

3) better comparisons between diodes can be made if one targets specific Currents rather than Voltages, discussion on this below.

A view of my current data reporting spreadsheet showing ALL input parameters as well as determined values of n, Is, and Rd.
Data input fields in blue: (target Id1 close to 1.0 uA and Id2 close to 0.5 uA).
Calculated fields in red and black.
Adjustment field "n" in green: (adjust value of n until the ratio of the two calculations = exactly 1).

Spreadsheet for calculation of diode n and Is							Tuggle Method from "Cal_n_Is.xls"						
Diode		n	Is	Rd	Temperature		Vd2	Id2	Vd1	Id1	Calc	Calc. Using	Calc. Using
under test			nA	k Ohm	dF	dC	V	uA	V	uA	ratio	Vd1 and Id2	Vd2 and Id1
HP 5082-2835		1.0713	12	2206	80.2	26.8	0.10215	0.4961	0.12215	1.0384	1.0000	4.124E-05	4.124E-05
1N5711	blue	1.0683	6	4541	80.0	26.7	0.12115	0.4917	0.13981	0.9757	1.0000	7.931E-05	7.931E-05

(Note: the engineers in the crowd will have noticed in the above report than I am carrying a degree of precision not justified by the level of accuracy in my measurements. The final results of n, Is, and Rd should be rounded off to not more than three significant figures.)
A copy of my spreadsheet can be downloaded here..

While measuring a good number of diodes, both Germanium, Schottky, Silicon, and even a few LED's, I noted that Tongue's recommended measurements at Vd = 0.039 and 0.055 V resulted in current reports spread out over more than two orders of magnitude. As the determination of Is and n is not unique but a function of Id, I felt uneasy by this method. While most Germanium diodes I have measured have a fairly narrow range of Forward Voltage drops, (Vf), that of Schottky's can range up to a tenth of a volt. This is the factor responsible for the huge range of measured Id values, see the following plot of I/V characteristics for some selected diodes:

Schottky Law with I*Rs Term

Current, I:

$$I = Is \{ \exp [q (V - I *Rs) / nkT] - 1 \} \text{ (implicit expression)}$$

Dynamic Conductance, dI / dV:

rearranging, let: $\exp [q (V - I\ Rs) / nkT] - ((I / Is) + 1) = 0 = F(V, I)$

then: $dI / dV = (- dF / dV) / (dF / dI)$

$(- dF / dV) = \exp [q (V - I\ Rs) / nkT] * (q / nkT)$, and

$(dF / dI) = \exp [q (V - I\ Rs) / nkT] * (-q\ Rs / nkT) - 1 / Is$

then,

$$dI / dV = 1 / \{ (nkT / q\ Is) * (\exp [-q (V - I\ Rs) / nkT] + Rs \}$$

Dynamic Resistance, 1 / (dI / dV):

$$Rd = dV / dI = (nkT / q\ Is) * (\exp [-q (V - I\ Rs) / nkT] + Rs$$

Curvature, d2I / dV2 :

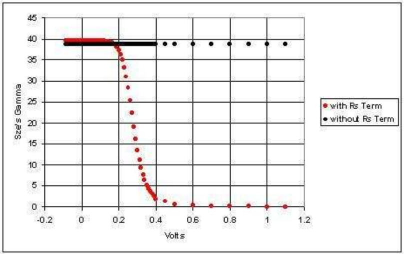
$$d2I / dV2 = d/dV [1 / \{ (nkT / q\ Is) * (\exp [-q (V - I\ Rs) / nkT] + Rs \}]$$

Let: $(nkT / q\ Is) * (\exp [-q (V - I\ Rs) / nkT] + Rs = U$

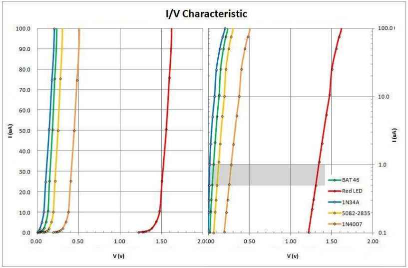
3. With increasing V, the slope (or conductance) itself increases initially, and then levels off at a constant value of 1 / Rs. (red curve)

4. Curvature, the rate-of-change in slope with increasing V, increases initially to a maximum level. After that it decreases and goes to zero as the straight-line portion of the curve is reached. (blue curve)

5. Sze's sensitivity figure, curvature over slope, starts out constant initially but decreases to zero as curvature goes to zero. (red curve in figure below -- compared to the black curve case with no Rs term)



6. Note that Sze's gamma takes a dive near the same voltage where curvature peaks. (??)



In order to get what I feel is a better comparison between diodes, I have decided to target not some pre-determined voltage, but rather target currents of Id = 0.5 and 1.0 uA. On the above plot on the right (with Log Id vs Vd scale) it will be clear that I am aiming at the same part of the characteristic curve regardless of Vf. Hopefully this will allow good comparison between diodes with rather different Vf, even as far as including Silicon and LED's in my mix. On va voir....

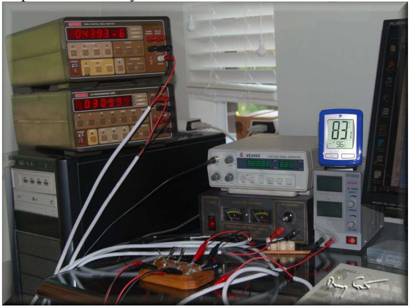
Shockley equation Id = Is*{exp[(qe/(n*k*T))*(V-I*Rs)]-1}
where:
n = ideality factor
Is = Saturation current in Amps
Id = Diode Current in Amps
Vd = Diode Voltage
k = boltzmann = 1.38E-23 J/K
T = temp K = 300
K = dC + 273.15 Kelvin
qe = electron charge = 1.609E-19 cmb

Often simplified to:
Id = Is*(exp(Vd/(0.0256789*n))-1)   [at room temperature]

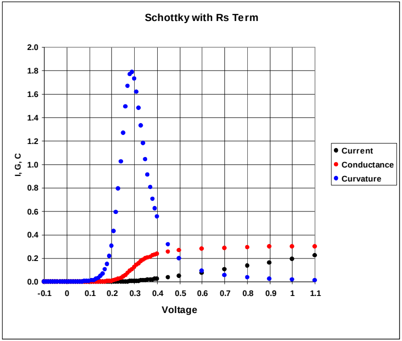
Diode Resistance Rd = VT * n / Is
Where:
T = 300K
VT = k*T/qe = 0.0256789
SO: Rd = k * T / qe * n / Is

I cannot express enough my indebtedness and thanks to Ben Tongue and Ken Khun for their great work and web documentation in matters Diode!

A photo of the current "bench" setup. Keithey 195A pico-ampmeter + Keithley 192 DMM bench meter.



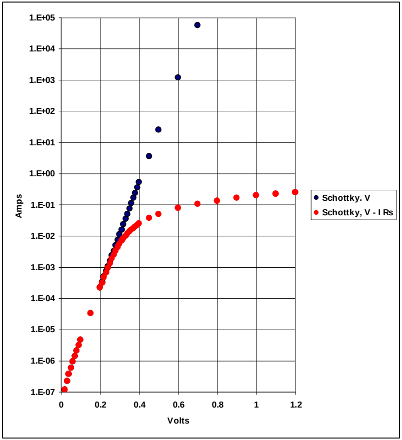
Kevin 09/2011



Diode: Is = 1e-7 A.; eta (n) = 1; Ro = 258 k-ohms; Rs = 3.2 ohms; T = 300 K.

1. At low voltages, current rises exponentially at first but then increases linearly as the I*Rs voltage term dominates. (black curve)

2. Initially resistance (1/slope) decreases exponentially from a high value. Then it becomes constant as barrier resistance R becomes vanishingly small and parasitic resistance Rs becomes the dominant resistance. (resistance curve not shown)



Schottky Behavior with Rs Term

Diode voltage-current behavior is modeled with the Schottky expression that includes the I*Rs term. Below are graphs of some of the resulting quantities.

Spreadsheet for calculation of diode n and Is (@1.0uA)

Diode		n	Is nA	Rd k Ohm
Germanium Diodes				
FO-215	ITT	1.11	175	163
FO-215	ITT	1.10	196	144
1N277	black	1.50	2293	17
1N277	black	1.60	2296	18
1N270	bonafide	1.28	915	36
1N270	blue	1.26	857	38
1N270	blue	1.66	2310	19
D18	russia	1.20	194	160
D18	russia	1.27	193	170
GAZ 51	Tesla	1.14	140	211
GAZ 51	Tesla	1.52	617	64
OA 5	Tesla	1.85	3384	14
OA 5	Tesla	1.48	1996	19
D9E	russia	1.42	2161	17
D9E	russia	1.54	2414	16
1N34A	bonafide	1.82	2153	22
1N34A	bonafide	1.25	1392	23
1N34A	green	1.57	1389	29
1N34A	green	1.30	1185	28
1N34A	green	1.32	1554	22
1N34A 37	orange	1.10	1458	19
1N34A 37	orange	1.36	1370	26
1N34A	red	1.34	833	42
1N34A	red	1.32	993	34
D310	russia	1.01	1215	21

Diode		n	Is	Rd
			nA	k Ohm
GD 402A	russia	1.55	970	41
GD 402A	russia	1.71	1360	33
UK A	russia	1.56	1565	26
UK A	russia	2.00	5766	9
UK B red	russia	1.64	2049	21
UK B red	russia	1.85	3777	13
UK C ora	russia	1.92	3462	14
UK C ora	russia	1.95	3354	15
UK D blk	russia	2.00	4901	10
UK E blk	russia	2.00	5862	9
UK F blk	russia	1.82	3523	13
UK G blue	russia	1.86	3123	15
Average		1.39	1334	54
Schottky Diodes				
HP 5082-2835		1.07	12	2206
1SS98		1.08	12	2302
1SS98		1.06	13	2188
1N5711	blue	1.07	6	4541
1N5711	blue	1.07	6	4816
1N60		1.09	178	158
1N60		1.10	161	176
BAT 46		1.12	141	204
BAT 46		1.18	172	177
1N34A ?	schottky	1.11	317	90
1N34A ?	schottky	1.24	482	66
1N5819		1.19	750	41

Effect of Parasitic Resistance Rs on Diode Current

The common Schottky expression predicts extraordinarily high currents for applied voltages above several hundred millivolts (black curve). For example, at the "knee" voltage, 0.7 volts, predicted current is on the order of 50,000 amps! Of course, this doesn't happen. When one includes the effect of series parasitic resistance Rs, replacing V by (V - I Rs), current is tamed to a more reasonable 100 milliamps in this example.

When I*Rs term is included, I appears on both sides of the Schottky expression and cannot be rearranged into a single I term. It must be solved for by trial-and-error. Excel's SOLVER function can be used, but still, each data point must be solved separately.

Diode: Is = 1e-7 A.; eta (n) = 1; Ro = 258 k-ohms; Rs = 3.2 ohms; T = 300 K.

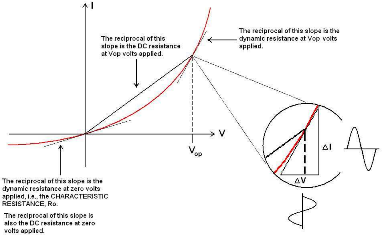
DC RESISTANCE (Shockley-Law):

$$RDC = V / I = (nkT / q*I) * \ln [(I / Is) + 1] + Rs$$
$$RDC = R + Rs$$

Dynamic resistance Rd is the impedance (resistive) offered by the diode to an alternating (RF) signal when the diode is biased by a DC current I. Rd is the inverse of the slope of the DC I versus V curve at voltage point V (or current I). The dynamic resistance at zero volts DC applied is the characteristic resistance. It is the resistance to be considered when matching the diode to, say, the tank circuit in a crystal set. See illustration below.

DYNAMIC RESISTANCE:

$$Rd = dV/dI = d/dI [(nkT /q) * \ln [(I / Is) + 1]+ I * Rs]$$
$$Rd = (nkT / q*I) * \ln [(I / Is) + 1]$$

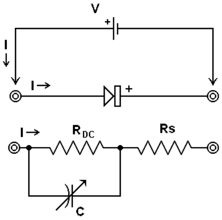


1N5819		1.14	807	36
Diode		n	Is	Rd
			nA	k Ohm
1SS16	NEC	6.24	14288	11
1SS16		2.27	4101	14
Average		1.13	276	1137
Silicon Diodes				
1N914		1.95	5	9293
1N914		2.02	7	7618
1N4148		2.00	6	8934
1N4148		2.00	6	9323
1N4007		1.51	1	64311
1N4007		1.51	1	77954
6A10		1.57	2	19275
6A10		1.54	2	20379
1N4736A	Zener	1.15	0	1.91E+09
1N4736A	Zener	1.14	0	2.50E+09
KB 130	russian	1.14	0	5.83E+08
KB 130	russian	1.16	0	4.36E+08
UK H	russian	1.94	8	6263
UK H	russian	1.77	4	12548
UK I	russian	1.78	0	260253
KD 401A	russian	1.51	0	232930
KD 401A	russian	1.38	0	1539619
D 220	russian	1.26	0	322405
D 220	russian	1.24	0	537707

D 223A	russian	1.20	0	10262661
Diode		n	Is	Rd
			nA	k Ohm
D 223A	russian	1.20	0	8565213
UK J	russian	1.46	4	9849
UK J	russian	1.40	2	20013
Average		1.52	2	2.37E+08
Light Emitting Diodes				
Red		2.15	0	1.70E+12
Red		2.07	0	5.52E+12
Amber		1.56	0	2.09E+17
Amber		1.62	0	6.46E+16
Yellow		1.59	0	8.03E+17
Yellow		1.38	0	2.87E+20
Green		2.05	0	4.23E+13
Green		2.07	0	4.34E+13
Water Green		3.27	0	4.85E+13
Water Green		1.93	0	5.79E+19
White-Gn		1.44	0	1.05E+20
White-Gn		3.04	0	6.88E+09
Average		2.01	0	3.76E+19

Diode Model
Notes provided to the editor by Mike Tuggle

The diode is modeled as follows:



If a DC voltage V is placed across the diode, current I flows through the diode.

RDC is the non-linear, Shockley-law DC resistance that varies with V and I. RDC is determined by diode-specific parameters Is and eta (n) in the exponential expression.

Rs is the fixed parasitic (bulk) resistance that stays constant with V.

C is the junction capacitance that can be ignored when considering just the direct current (DC) behavior of the diode.

CURRENT:

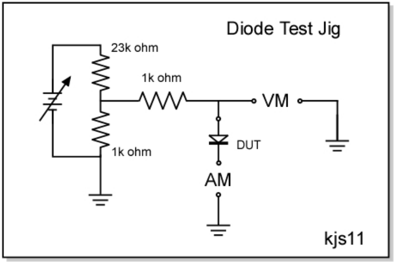
$$I = Is \{ \exp [(V - I*Rs) / (nkT)] - 1 \}$$

VOLTAGE:

$$V = (nkT / q) * \ln [(I / Is) + 1] + I * Rs$$

Diode Test
Kevin Smith
http://www.lessmiths.com/~kjsmith/crystal/dtest.shtml

So, what is the best diode out there to use? Do crystals always show poor sensitivity compared to germanium? What about Fleming's marvelous device, the vacuum diode? I start by using a test jig, the design shamelessly modified from Dan Petersen in the Nov 2009 Crystal Set Society Newsletter, (if you aren't a member, go sign up to be.. NOW):

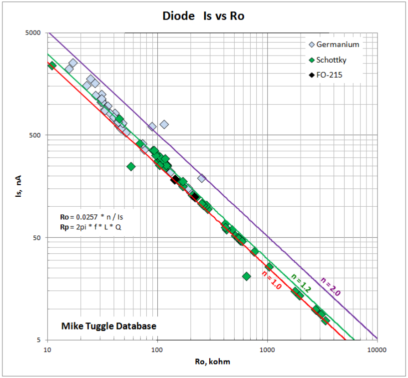


This is a very simple circuit allowing power input, attaching a diode for testing, and sampling both the current and voltage across the diode under test, (DUT). I made a simple modification by adding a voltage divider, about 20:1 at the power input. The power supply can be connected either at the front across the divider, or after the divider in front of the 100k current limiting resistor for higher voltage readings. I get the input voltage from a modest 18V, 2A regulated DC power supply. This unit has both volt and amp controls with digital readouts, nice to use and doesn't take up much room on my desk. Here is a case where I recommend you look at getting a new unit. I looked long and hard on ebay, but found few bargains and few units with the same features. Still, the supply regulator was never designed to adjust the output to precise small values making it somewhat difficult/tedious to adjust. Adding a small rheostat to the test jig may aid for making fine adjustments on the voltage. Another piece of equipment you will need for this test is an amp meter. At first I used a nice small digital meter I found new on ebay for $15 shipping included. It reads amps to a lowest range of 200uA, more than enough precision for the task. For that price get two. Later as I got to making more precision measurements for the determination of diode n and Is I acquired a Keithley 195A bench volt/amp meter with very high input impedance. The DVM is, of course, my trusty Keithley workhorse.

I have received occasional criticism for the simple test jig with the suggestion that proper measurements cannot be taken with the jig as-is. I note that, while the test setup is simple, the instrumentation is high quality and performs the job well. I have had the opportunity to collaborate with Mike Tuggle, comparing our respective measurements. Even my early characteristic curves taken with the small amp meter compare well and provide a good view of the diode and determination of the diode Rs series resistance. The detail n and Is measurements presented here essentially match, though at somewhat less precision, those data from Mike. The reader should have some assurance that the data and measurements found on this page are vetted and reliable.

Easy protocol as follows:

1) Set up the test with the diode and meters clipped to the jig.



I absolutely agree, your data set on diodes is preeminent for its detail and variety of diodes covered. I'll attach a summary index of all the diodes I've measured (ed note: summary chart above). Each diode has a full multipoint analysis and I-V plot, if you're interested in any of those. Quality varies, as the methodology has improved over the years.

Best regards, -Mike-

Hi, Kevin.

Coincidentally, I have been thinking about the measurement of diode I-V's at higher voltage levels. For an 0.4 volt drop across the 1N34A that we've been considering, the Shockley law current through the diode will be some 312 mA.

To measure that with a 50 k-ohm current sense resistor, we would need to supply 15,600.4 volts -- 0.4 volts across the diode; 15,600 volts across the current sense resistor. Obviously not practical, for my setup anyway.

A much smaller current sense resistor is needed. I'd say about 1 ohm. Then there would be 0.4 volts across the diode and 0.312 volts across the current sense resistor -- or 0.712 volts total. That could be done. Whatever current sense resistor is used, its value needs to be known accurately.

Under these conditions, the DC resistance of the 1N34A is 0.4 volts/0.312 amps = 1.28 ohms. The bulk (parasitic) resistance of the diode would need to be considered. I would suggest multiple measurements regressed by Excel's SOLVER, using the Shockley expression with bulk resistance parameter. Some notes are attached. (p 153).

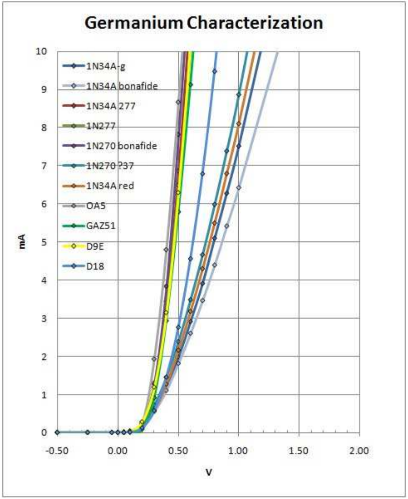
2) Adjust the power supply until the reading on the DVM meter is where you plan to measure. Adjust the voltage (DVM readout) in steps from 0 to 1.0 Volt in 0.1 V increments and record the current (Amps) in a spreadsheet.
3) Repeat at each tenth volt recording while the spreadsheet makes a sweet graph.

Test Results

First, I offer a quick look at my diode modeling setup. I am currently upping the ante on resolution in order to determine diode parameters Is and n, hoping for success soon! Setup and protocol..

Over a period of a few weeks I ran I-V characterization tests on a good number and variety of diodes, crystals and my two tubes to see how very thing stacks up. The following presents the resulting curves, diode photos, and some puzzling questions/conclusions I churned up in the process. I start off with the realization that when one orders diodes, it is by no means certain what one will get. This seems especially true for the supposedly ubiquitous 1N34A. Unless you see the part number actually on the diode, you probably need to test it to know what it really is. So, lets take a look!

Germanium Diodes



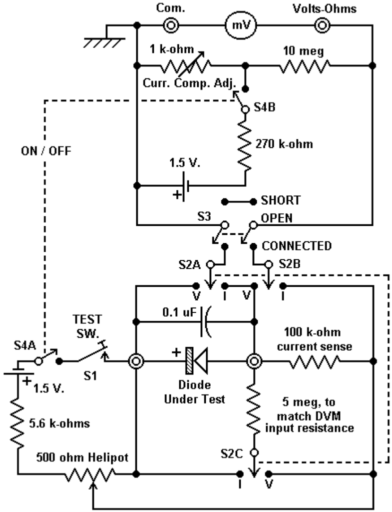
I can get the full 100 mV. With low resistance diodes, the attainable voltage is proportionately less.

My measurement sets typically consist of 20 (10 + and 10 -) V-I measurements. I then regress these measurements against the Shockley equation. (I minimize, using Excel's SOLVER, the sum of squared differences between measured currents and Shockley model currents, by adjusting eta (n) and Is.) Parasitic (bulk) resistance is neglected.

I may be able to get higher currents through a 1N34 by going to a 1.5 volt C cell and shunting across the 5.6 k-ohm voltage divider resistor. I don't want to mess with the 100 k-ohm current sense resistor -- it gets too complicated.

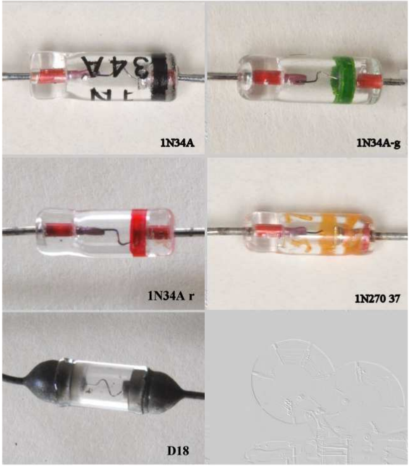
Here's a shot of my test stand:

Mike

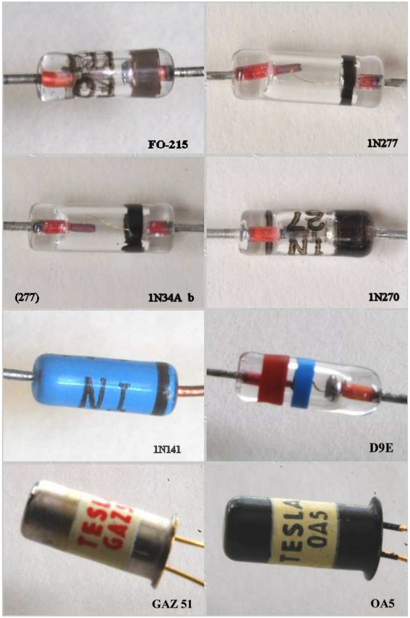


Test voltage is supplied by a single 1.5 volt AA cell. Design philosophy was to cover voltages (and currents) experienced by diodes in crystal sets. This setup can supply over 100 mV across the diode. The maximum voltage across a diode depends upon the resistance of that diode. With silicon diodes,

1N34A series

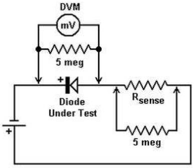


1N270-277 series

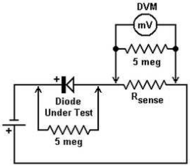


Diode Test Apparatus
Mike Tuggle

Hi, Kevin. My basic diode test circuit is:



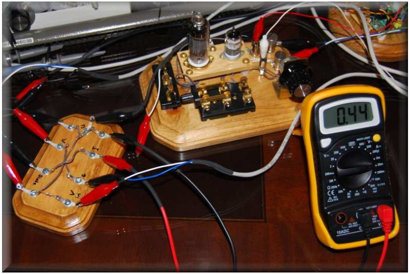
Basic Diode Voltage Measurement



Basic Diode Current Measurement

Diode voltage and current measured sequentially by the same DVM in voltage mode. The 5 meg resistor compensates for the 5 meg DVM input resistance. Here's the complete measurement circuit designed with much help from Ben Tongue:

Photo of the lab with my diode test jig, amp meter, and diode vacuum tubes under test.



Here we see a set of Germanium diodes and photos of the diodes under test. Immediately you should notice two sets of curves, what I call the "1N34A" set and the "1N270-277" set. In this analysis, the second set appears clearly superior in sensitivity and deservedly so. In the photos it is clear that one bunch of diodes I purchased as 1N34A were in fact better 1N277, the black band and thin gold contact wire as well as common curve unite these. "True" 1N34A's have a more robust contact wire. I also purchased a 1N270 and received pretty orange diodes with the number "37" stamped on it, a fairly robust contact wire and a curve that most closely resembles a 1N34A, go figure. Mercifully, I DID manage to get some bona-fide 1N34A's with the part number on the diode, as well as a single bona-fide ITT 1N270 diode with the part number. These are my base for comparison.

In addition to the standard Germanium Diodes tested above, I also have latched my hands on a few "other" germanium diodes for the fun of testing. These include two vintage Russian types, a D9E (in the 270 class) and a D18 (transitional between 270 and 1N34A). Additionally, I got a few FUZZ diodes popular, I suspect, with the guitar gadget crowd. These are larger packages in metal cases so I cannot see how the internal contact is made other that they are stated to be gold bonded point contact type. Clearly these are highly sensitive diodes in my 270 class, the OA5 looking best of the lot. Interestingly, I recently purchased from good Mr. Peebles a few of his "Holy Grail" ITT FO-215 diodes. My measured characteristic for one of these is a dead laydown on the gold-bonded OA5. From the photos, these are in different packaging with the FO-215 in a traditional glass casing. What makes the FO-215 so great is the fact that its resistance Rd is an interesting 150k ohm or so which matches very well with typical moderate Q tank circuits. Interestingly, I find that an old Russian D18 diode has very similar characteristics to the

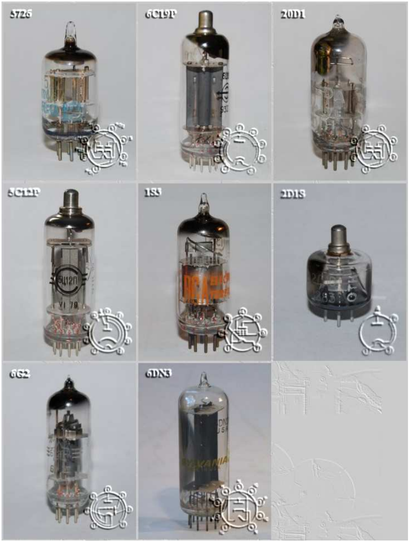
FO-215 and so much also be placed into the "Holy Grail" class of germanium diode. Take your pick!

Spreadsheet for calculation of Germanium diode n and Is (modified from Mike Tuggle's spreadsheet)

Spreadsheet for calculation of diode n and Is					Tuggle Method from "Cal_n_Is.xls"					
Diode		n	Is	Rd	Temperature		Vd2	Id2	Vd1	Id1
under test			nA	k Ohm	dF	dC	V	uA	V	uA
FO-215	ITT	1.0993	196	144	80.7	27.1	0.03574	0.49730	0.05123	1.00250
FO-215	ITT	1.1050	175	163	80.9	27.2	0.03882	0.50990	0.05735	1.13850
1N270	blue	1.6610	2310	19	82.4	28.0	0.00835	0.49630	0.01635	1.07150
1N270	bonafide	1.2790	915	36	82.4	28.0	0.01432	0.49660	0.02437	0.99860
1N270	blue	1.2590	857	38	82.5	28.1	0.01505	0.50440	0.02829	1.18850
1N277	black	1.4990	2293	17	81.5	27.5	0.00765	0.50180	0.01402	1.00250
1N277	black	1.6042	2296	18	81.6	27.6	0.00821	0.50400	0.01639	1.11620
1N141	blue	1.1795	726	42	85.1	29.5	0.01642	0.51540	0.03075	1.25650
1N141	blue	1.2570	1402	23	84.5	29.2	0.01024	0.51750	0.01795	1.02980
D310	russia	1.0059	1215	21	83.6	28.7	0.00915	0.51170	0.01695	1.11490
D9E	russia	1.5350	2414	16	81.1	27.3	0.00748	0.50250	0.01522	1.13280
D9E	russia	1.4200	2161	17	80.6	27.0	0.00755	0.49570	0.01402	1.01000
GAZ 51	Tesla	1.5152	617	64	83.1	28.4	0.02356	0.50850	0.03782	1.00240
GAZ 51	Tesla	1.1379	140	211	82.9	28.3	0.04491	0.50280	0.06275	1.03850
OA 5	Tesla	1.8490	3384	14	83.4	28.6	0.00664	0.50370	0.01255	1.01480
OA 5	Tesla	1.4810	1996	19	83.3	28.5	0.00865	0.50550	0.01645	1.07020
1N34A	bonafide	1.8220	2153	22	81.6	27.6	0.00995	0.50780	0.01911	1.08050
1N34A	green	1.3210	1554	22	82.9	28.3	0.00945	0.49530	0.01695	0.99850
1N34A	bonafide	1.2500	1392	23	82.2	27.9	0.00995	0.50250	0.02014	1.20570
1N34A	green	1.3030	1185	28	83.1	28.4	0.01175	0.49420	0.02135	1.04750
1N34A	green	1.5740	1389	29	82.7	28.2	0.01237	0.49350	0.02327	1.07190
1N34A	red	1.3220	993	34	83.6	28.7	0.01401	0.50250	0.02475	1.05390
1N34A	red	1.3430	833	42	83.3	28.5	0.01675	0.51570	0.02765	1.01250
1N34A 37	orange	1.0955	1458	19	83.3	28.5	0.00861	0.51740	0.01495	1.01250
1N34A 37	orange	1.3550	1370	26	83.6	28.7	0.01091	0.49990	0.01925	1.00190
D18	russia	1.2040	194	160	82.5	28.1	0.04004	0.50910	0.05913	1.10500
D18	russia	1.2706	193	170	82.9	28.3	0.04215	0.50250	0.06415	1.16550
GD 402A	russia	1.7080	1360	33	83.6	28.7	0.01365	0.49200	0.02785	1.19350
GD 402A	russia	1.5535	970	41	83.4	28.6	0.01719	0.51780	0.03185	1.17270

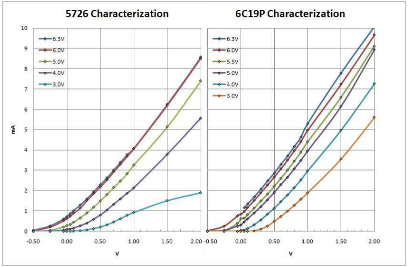
Averages: n = 1.38 Is = 1315 nA Rd = 52 kOhm

The above spreadsheet is based on measurements of the diodes shown above. For the determination of Is and n, I chose at random two examples from my collection of various diodes, (all germanium in this case) and measured via a modified version of the methodology outlined by Ben Tongue and Mike Tuggle. For this work I have chosen to set Id2 about 0.5uA and Id1 about 1.0uA and then read the needed voltage. This is the inverse of the normal method but the justification is that for



Small currents will continue to flow even with a negative bias to the plate, (Contact potential).

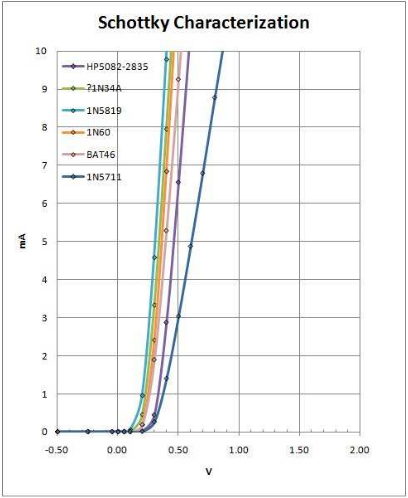
In order to explore this idea, I ran a series of tests with the tubes running at lower operating voltages (effectively diminishing the space charge thus lowering the perveance) as shown:



Here you see the impact of lower operating voltages. With the tube operating between 3 and 4 volts the characteristic curves begin to pass through the origin as in regular solid-state diodes and crystals. It is precisely in this voltage range that I found the radio sensitivity, as measured by loudness to my ear, to be most pronounced. It appears that the diode characteristic needs to pass through the origin of the I-V graph for highest sensitivity. This takes me to the 20D1 tube which operates at 9v but only 200mA. This will still take a power supply to use, but the good thing is that the characteristic curve passes through the origin. Running at its design parameters the tube looks a lot like the 6C19P at 4v.

any diode regardless of forward voltage drop the measurement is made at the same part of the LOG I vs V characteristic. This allows comparison between all diodes. I have included the actual room temperature in the calculations in order to eliminate this variable as a source of doubt or error. All the measured parameters as well as the determined results are presented in order to facilitate repeatability.

Schottky Diodes

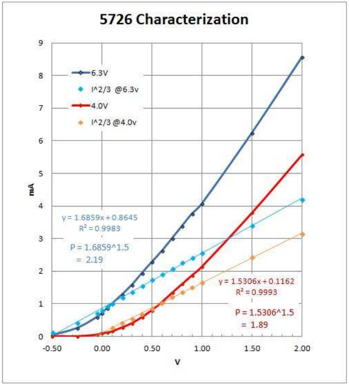


The perveance numbers above range from 0.02 to over 2 mA/V^3/2 (with >2 being good for small-signal detection) and clearly shows my choice of the 6C19P to be an excellent one. I am surprised to see a huge 5.4 for the 6DN3 diode, a color television damping diode. It takes a Novar 9-pin socket and is a large tube so doing more with this tube may have to wait a bit.

Calvert's 2001 published data

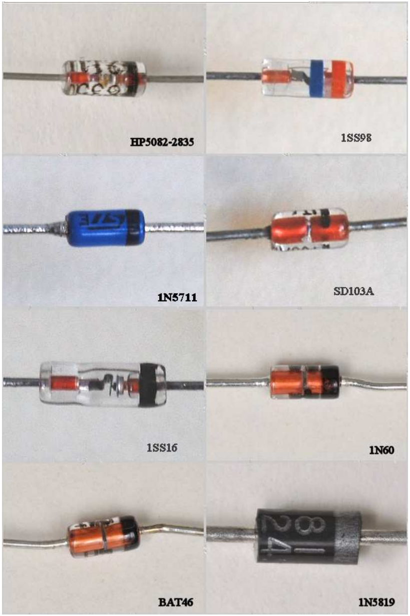
Tube	Perveance
6AL5	2.42
6H6	0.50
7Y4	0.58
2X2A	0.017
6V3-A	2.3
6AX4GT	1.42
6AV6	0.085
1A3	0.075

The high-perveance diode tube types I tested show an interesting property in that their characteristic curves do not pass through the origin of the graph. These curves are wholly acceptable as crystal radio detectors although they will require power to operate. What I have noticed with such tubes is that when I turn them off after using them with a radio, on cooling they go through a period of much increased sensitivity (very loud) before fading to nothing. It is as though running them at the full 6.3V lowers their full potential as crystal radio rectifiers. The reason is in the high perveance of the tubes, the anode is already proximate to the space charge before any plate voltage is applied. Effectively, while the plate (anode) "Zero" voltage is measured with respect to ground, the plate itself is still positive with respect to the cathodic space charge.



With my measurements calculating the perveance is practical and quickly done, results as follows:

Tube	Perveance	R2
6C19P	2.54	0.997
5726	2.19	0.998
6DN3	5.40	0.997
20D1	2.02	0.985
5C12P	0.32	0.999
1S5	0.063	0.999
2D1S	0.155	0.975
6G2	0.050	0.999



Schottky diodes are very sensitive diodes that work excellently in crystal radios. Their construction and theory are different and I confess to not fully understanding these components. Still, from the characteristic curves, they are excellent! Again, note that you dont always get what you bargain for. Here I found what was supposed to be 1N34A's to be some sort of Schottky of unknown pedigree. The 1N5819, while having the lowest forward voltage drop, has a rather very high Junction Capacitance and will not perform well. Posts on the RaidoBoard Crystal Radio Forum however highly recommend the 1N5711 for crystal sets although the characteristic curve dosnt look that fabulous. Many web pages out there on Shottky diodes, I recommend you do your homework. Of all the schottky's, I note that Ben Tongue recommends most highly the HP5083-2835. The high resistance Rd makes them useful for DX sets with very very high Q tanks. Even so the diodes need to be paralleled with up to 4 or 5 diodes to match correctly the Rd with the tank Rp. I have found these somewhat hard to find and expensive, especially when one requires using several in parallel. I recently measured a few 1SS98 diodes and I discover them to have characteristics extermely similar to the HP and I feel they deserve more attention. Sadly, on ebay they seem to be just as difficult to find and expensive. No free ride!

Spreadsheet below for calculation of Schottky diode n and Is (modified from Mike Tuggle's spreadsheet)

Langmuir-Child Law, the steepness of the I/V characteristic is largely due to the geometry of the tube elements and the volume of electron space-charge between them. JB Calvert's Theory of Vacuum Tubes informs one that this is the property called Perveance in a vacuum diode. Measuring this requires plotting I raised to the 2/3 power against V and taking the slope of the best-fit line through the data. This slope raised to the 3/2 power is the perveance. The following chart illustrates graphically the relation bewtween the tube characteristic and perveance. For the 5726 tube I plot the characteristic in dark blue for the tube heated with the design 6.3 volts. In light blue I have plotted the measurements and raised them to the 2/3 power. One sees immediately that the new curve is linear. The slope of the curve raised to the 3/2 power yields the perveance factor for the tube.
(Note in addition that when running the tube at a more sensitive point with 4 volts only on the heater the perveance is slightly less than spec.).

Vacuum Diodes

When I decided early on to construct a radio based on the vacuum diode (see my Fleming Radio section), I had to find a suitable tube for the project. That began with an extensive search through the datasheets checking vacuum diode properties and reviewing the characteristic curves for a large number of diode tubes. In the above figure I plot the characteristics as best I can on a commpn plot for comparison. I was hoping to find a good candidate with sensitive characteristics and low energy consumption. As you will find, such a beast did not exist. Following this research began a period ot purchase and testing.

I have tested seven different tube types including 6.3V rectifier diodes that take a lot of power to run and are not really suitable for battery use, a 9v dual diode tube (20D1), a pentode/diode tube (1S5) that is designed to run on a battery at 1.5V and 50mA, and a couple miscellaneous but cute tubes. In testing the 1S5, I found the filament never glowed incandescant at 1.5V and, on measuring, was barely sensitive to anything. I cannot imagine this tube would make much of a diode for crystal radio use. I tested different manufacture 1S5 tubes from two different suppliers. No dice. Finally, I tried to push the tube to operate at higher-than-specified voltages. At 4.3V the I-V curve was still very flat, but shifted slightly higher, crossing the Amp axis at about 0.4 mA, (see graph). At 5.5V one of my tubes gave up the ghost and I didnt do more. I pretty much rule the 1S5 out for my crystal radio work. Looks like i'll be needing the power supply when I get around to building/operating that Fleming Radio.

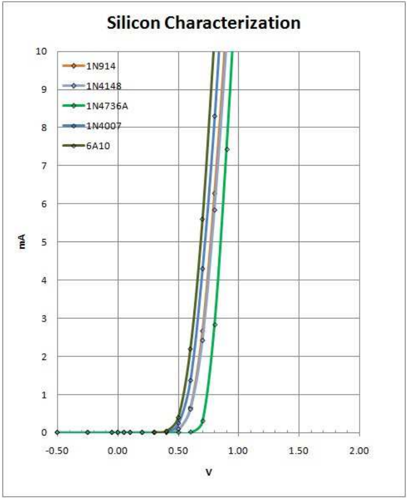
A question remains as to what in fact is different about these tubes to give such different characteristics. While the I/V characteristic of most vacuum diodes pretty much follow the

Spreadsheet for calculation of diode n and Is							Tuggle Method from "Cal_n_Is.xls"			
Diode		n	Is	Rd	Temperature		Vd2	Id2	Vd1	Id1
under test			nA	k Ohm	dF	dC	V	uA	V	uA
HP 5082-2835		1.0713	12	2206	80.2	26.8	0.10215	0.4961	0.12215	1.0384
1SS98		1.0626	13	2188	83.8	28.8	0.10245	0.5081	0.12085	1.0036
1SS98		1.0753	12	2302	83.8	28.8	0.10430	0.5001	0.12383	1.0208
1N5711	blue	1.0683	6	4541	80.0	26.7	0.12115	0.4917	0.13981	0.9757
1N5711	blue	1.0733	6	4816	80.2	26.8	0.12323	0.4917	0.14261	0.9979
SD103A	50pF	1.0443	112	241	83.3	28.5	0.04611	0.5053	0.06185	0.9934
SD103A	50pF	1.0435	109	247	83.4	28.6	0.04648	0.5021	0.06305	1.0201
1N60		1.0884	178	158	80.4	26.9	0.03743	0.4988	0.05311	1.0065
1N60		1.1040	161	176	80.0	26.7	0.03989	0.4955	0.05729	1.0508
BAT 46		1.1185	141	204	80.0	26.7	0.04304	0.4878	0.06032	1.0055
BAT 46		1.1839	172	177	80.7	27.1	0.04173	0.5048	0.05839	0.9972
1N34A ?	schottky	1.1104	317	90	80.4	26.9	0.02710	0.5019	0.04068	1.0003
1N34A ?	schottky	1.2359	482	66	80.4	26.9	0.02269	0.5016	0.03615	1.0197
1N5819	100pF	1.1881	750	41	80.6	27.0	0.01551	0.4957	0.02586	0.9975
1N5819	100pF	1.1425	807	36	80.7	27.1	0.01453	0.5155	0.02541	1.1074
1SS16	NEC	6.2400	14288	11	81.3	27.4	0.00564	0.5099	0.01165	1.0733
1SS16	NEC	2.2700	4101	14	81.5	27.5	0.00684	0.5083	0.01385	1.0947

Averages: n = 1.48 Is = 1275 nA Rd = 1030 kOhm

The above spreadsheet is based on measurements of the diodes shown above. For the determination of Is and n, I chose at random two examples from my collection of various diodes, (all schottky in this case) and measured via a modified version of the methodology outlined by Ben Tongue and Mike Tuggle. For this work I have chosen to set Id2 about 0.5uA and Id1 about 1.0uA and then read the needed voltage. This is the inverse of the normal method but the justification is that for any diode regardless of forward voltage drop the measurement is made at the same part of the LOG I vs V characteristic. This allows comparison between all diodes. I have included the actual room temperature in the calculations in order to eliminate this variable as a source of doubt or error. All the measured parameters as well as the determined results are presented in order to facilitate repeatability.

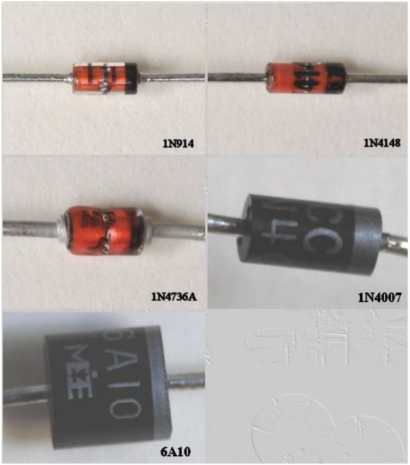
Silicon Diodes



In the photo I indicate groupings based on an easy measure of performance. I note the current in milliamps for each set where the plate voltage is set at 0.5V. The greater the current the better are your chances to get a sensitive crystal, assuming Ohm's Law is not followed! Crystals in the "dead zone" on the left will have their woods metal re-melted for new detector crystals and the bad ones tossed, its tough love for crystals. I find that easily half the potted crystals I make are tossed this way, and only a few can be considered superlative.

Galena" samples (Tintic Utah, or Leadville Colorado) there are numerous hot spots under virtually every place I touch the probe, but in general the sensitivity is good to so so. These crystals are very kind to work with in terms of finding spots and avoiding frustration. Mirror galena on the other hand may have quality hot spots, but any hot spots at all are rare and frustratingly difficult to locate. Here my Philmore detector crystal shines with an almost ideal "Galena" response. To chase down this rabbit I purchased some lovely mirror galena from Sweetwater Missouri. I broke off a few appropriate-size chunks to pot in woods metal and test. At first I was very excited with the high currents I was seeing at moderate voltages. Figured I had struck gold. When these crystals failed miserably to rectify anything in my radios, I re-measured things in both forward and reverse directions. These crystals obey Ohm's law and act like typical resistors, not suitable for radio work at all.

For my pyrite crystals the work has been especially tedious and frustrating. With one of the crystals one "hot spot" alternated, entirely on its own, between hot and bad while I was making the measurements. I would start over and over, sometimes getting interesting readings then suddenly it would drop to low values and I'd start over, back and forth. I present this data as best as I have measured, and I don't intend to go back! You see at least one of the crystals, my "China 1", (from a lead/silver mine in Hunan) gave a sweet classic-looking curve. More to the point, "ideal" curves for natural minerals, are difficult to come by. Most crystals you use will be less than ideal. The good news is that, while listening to your set, poking about for a good spot is far easier than what I have gone through to produce these curves. Your ear will take care of you!



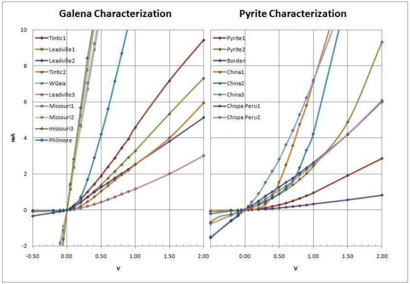
Silicon diodes have good characteristics, but an unacceptably high forward voltage drop making them a very poor choice for crystal radio unless used with bias. The 1N4736A is a Zenner diode.

Spreadsheet for calculation of Silicon diode n and Is (modified from Mike Tuggle's spreadsheet)

Spreadsheet for calculation of diode n and Is							Tuggle Method from "Cal_n_Is.xls"			
Diode		n	Is	Rd	Temperature		Vd2	Id2	Vd1	Id1
under test			nA	k Ohm	dF	dC	V	uA	V	uA
1N914		1.9510	5.4	9293	82.0	27.8	0.22733	0.4895	0.26229	0.9855
1N914		2.0236	6.9	7618	82.9	28.3	0.22354	0.4857	0.26423	1.0653
1N4148		1.9950	5.8	8934	83.1	28.4	0.23194	0.5117	0.27037	1.0841
1N4148		2.0041	5.6	9323	83.3	28.5	0.23177	0.4803	0.27346	1.0801
1N4007		1.5139	0.6	64311	83.3	28.5	0.26320	0.5041	0.29137	1.0355
1N4007		1.5140	0.5	77954	83.3	28.5	0.26951	0.4885	0.29807	1.0133
6A10		1.5680	2.1	19275	83.3	28.5	0.22129	0.4902	0.25023	1.0026
6A10		1.5368	2.0	20379	83.3	28.5	0.22021	0.4942	0.24764	0.9871
1N4736A	Zener	1.1505	0.00002	1.9E+09	83.4	28.6	0.51439	0.4961	0.53702	1.0609
1N4736A	Zener	1.1365	0.00001	2.5E+09	83.4	28.6	0.51663	0.5017	0.53648	0.9853
KB 130	russian	1.1403	0.00005	5.8E+08	83.4	28.6	0.47531	0.5012	0.49587	1.0061
KB 130	russian	1.1649	0.00007	4.4E+08	83.4	28.6	0.47567	0.4929	0.49746	1.0155
UK H	russian	1.9443	8	6263	83.6	28.7	0.20333	0.4486	0.24405	1.0175
UK H	russian	1.7727	4	12548	83.1	28.4	0.22717	0.5148	0.25707	0.9916

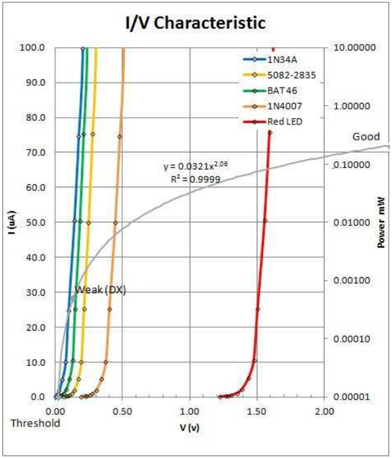
Averages: $n = 1.52$ $Is = 2$ nA $Rd = 2.4E8$ kOhm

The above spreadsheet is based on measurements of the diodes shown above. For the determination of Is and n, I chose at random two examples from my collection of various diodes, (all silicon in this case) and measured with a modified methodology to that outlined by Ben Tongue and Mike Tuggle. In this case, with the radically different forward voltage drop of these diodes from Germanium or Schottky diodes, I have kept the Id values constant (about Id2=0.5uA and Id1=1.0uA) and varied Vd. I have included the actual room temperature in the calculations in order to eliminate this variable as a source of doubt or error. All the measured parameters as well as the determined results are presented in order to facilitate repeatability.



The above curves demonstrate the wide variation in properties and qualities that can be found in natural crystals of galena or pyrite, the two most common and best quality natural stones. In each crystal test I first poked around the crystal for some time to 1) determine a typical sensitivity for the crystal in question and 2) to locate the best hot spot with which to test. This turns out to be a non-trivial exercise on the diode test setup. In a crystal radio one need only listen for the loudest spot. With the test setup one needs test both the forward and reverse current in order to determine if the whisker is on a hot spot or not. Very tedious work! (In retrospect, if I were starting over making a test jig, I would definately add a DPDT switch to readily change between forward and reverse current measurements. I'd probably toss in a rheostat for fine adjustments as well).

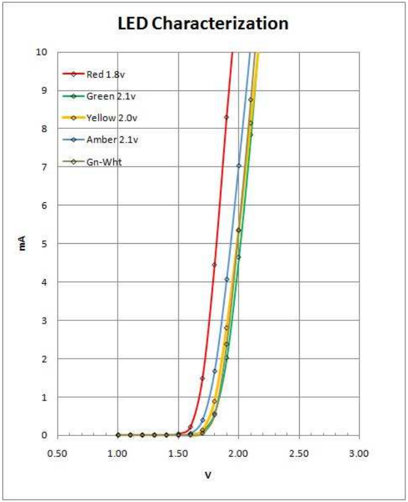
For many crystals there are limited number of possible hot spots, but these may be hot indeed. For most of my "Steel

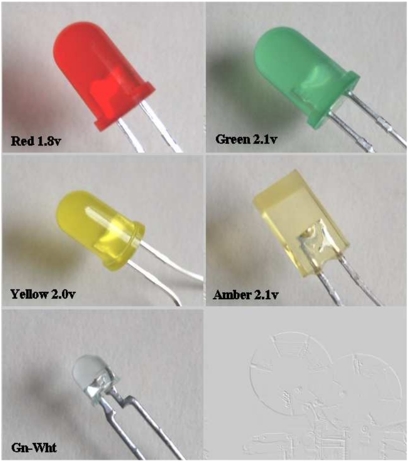


This plot at left takes some actual diode I/V measurements rather than the models presented below and plots them along with the above RF voltage across the LC circuit versus RF power in mW. Here one sees that pushing for very very low Vf (via high Is / low n combinations) pushes the limit of detectable signal power from the antenna. At such minute power levels I imagine the reverse leakage current becomes significant and probably offsets the low operating point gains sought.

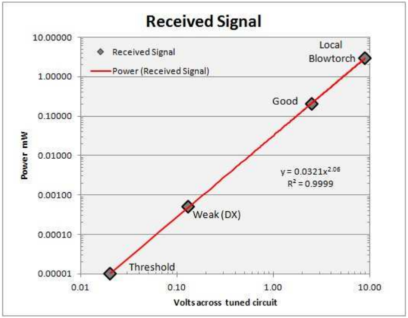
Crystal Galena and Pyrite Diodes

Light Emitting Diodes





As long as I am measuring various and sundry diodes, I figure I ought to include that most ubiquitous of modern diode, the LED. Found everywhere, these diodes are rapidly becoming the low-energy light source of choice for many lighting applications. I have read occasionally of someone asking whether these ought to be useful for radio applications as well. To this question the answer is generally a resounding "No!". The turn-on voltage is waaay above any reasonable value expected to be delivered by an antenna to a crystal set. Still,



The above is an interesting plot taken from the text discussion in "Crystal Set Analysis" by Berthold Bosch. It presents received signal strength across the tuned circuit for various scenarios from threshold audibility to local blowtorch. This plot should be considered a single example specific to his location and antenna/ground system. In the text he describes his antenna an an inverted L 43m long (140') and 10m high (32'), an excellent antenna most of us do not have the real estate to erect, but offsetting this is a poor ground with Rg = 210ohms. Given the offsetting conditions I would imagine this is a good generic example of what received signal voltage and power levels presented across the LC circuit to the diode are likely to be. Bosch cited 40nV as the threshold of audible detectability (impedance matched conditions with 16kohm RF impedance / 4kohm DC phones) and for this plot I pushed it back to 20nV as it gave a superior regression fit.

Additionally I have plotted RF signal Rs as orange circles from published coil unloaded Q (see my section on Coil Q). To convert from Qu to Ql I have made the assumption that loaded Q is about 1/5 unloaded Q for low-end sets while in performance sets it may approach 1/2.5 the unloaded Q. This is only an estimate then, but starting from actual data. To make the conversion to Rs I simply plugged the data into the standard equation R = 2pi*f*L*Q using a scaling factor to divide down Qu to Ql from high Q (1400/2.5) to low Q (100/5). This provides a nice visual display of the expected range for Rs in many crystal sets. I plot the Rs against the value of Ql used. Note that these data are for the case at about 1MHz and inductances between 200 and 300uH, as published.

Matching the diode to the tank is a matter of finding a diode with both sensitive qualities (low n) and an Rd close to the Rs presented to the diode (as per Ben Tongue's suggestions for Peak and Square Law detection). Those diodes with Is's in the 100 - 250 nA (100-280 kohm for n=1.1) or so range have the possibility to match while connected to the top of the tank coil without the use of Q-lowering taps, while untapped matching to high-end high-performance big-Litz baskets will require diodes with Is's in the 35 - 60 nA (500-800 kohm for n=1.1) range. Most "typical" germanium diodes have high Is values (>500, >60kohm) and will require a tap. Diodes such as FO-215, BAT 46, 1N60 and GAZ51 can be matched to many tanks without taps. For tanks with high quality Litz coils one will be using HP 5082-2835 or 1SS98 diodes, generally 3 to 4 in parallel to lower the Rd to the desired range. (Curiously, I have not found any diodes with Rd values in the 300 to 4500 kohm range.) Note that both resistance and impedance are frequency-specific so a match at one frequency will not remain matched across the broadcast band.

this simple answer avoids the actual question, what in fact does the characteristic curve of a LED really look like? Where is the turn-on voltage with respect to the published junction voltage, (assuming you can find that).

To provide just such a look I visited my local electronics store and bought a small handful of LED's, most with the junction voltage listed and took them home to measure. Typical LED junction voltages seem to range from about 1.8v to 2.1v or more. The turn-on voltages look closer to 1.6v-1.7v. Anyone used to working with Carborundum crystals or silicon diodes will be used to biasing their rectifier to get good sensitivity. These LED's once on have a very sharp rise and with a proper bias should work quite well as detector diodes. As a bonus you will get a sweet glow as well. OK, not as cool as the glow of a vacuum diode, but certainly more sensitive!

Spreadsheet for calculation of Light Emitting Diode n and Is (modified from Mike Tuggle's spreadsheet)

Spreadsheet for calculation of diode n and Is						Tuggle Method from "Cal_n_Is.xls"			
Diode	n	Is	Rd	Temperature		Vd2	Id2	Vd1	Id1
under test		nA	k Ohm	dF	dC	V	uA	V	uA
Red	2.1547	0.00000	1.7E+12	80.7	27.1	1.30114	0.4995	1.33895	0.9875
Red	2.0720	0.00000	5.5E+12	80.7	27.1	1.31649	0.5038	1.35690	1.0745
Amber	1.5627	0.00000	2.1E+17	80.4	26.9	1.42768	0.5046	1.45571	1.0131
Amber	1.6163	0.00000	6.5E+16	80.2	26.8	1.42525	0.4963	1.45532	1.0229
Yellow	1.5860	0.00000	8.0E+17	80.4	26.9	1.50347	0.5065	1.53292	1.0422
Yellow	1.3763	0.00000	2.9E+20	80.7	27.1	1.51826	0.5003	1.54291	1.0031
Green	2.0490	0.00000	4.2E+13	80.7	27.1	1.41068	0.5116	1.44647	1.0082
Green	2.0650	0.00000	4.3E+13	80.4	26.9	1.42153	0.5075	1.45762	1.0009
Water Green	3.2709	0.00000	4.8E+13	80.4	26.9	2.22240	0.5085	2.27970	1.0045
Water Green	1.9273	0.00000	5.8E+19	80.9	27.2	2.03170	0.5099	2.06510	0.9993
White-Gn	1.4366	0.00000	1.1E+20	80.9	27.2	1.54780	0.5143	1.57610	1.1050
White-Gn	3.0358	0.00001	6.9E+09	80.7	27.1	1.37740	0.5111	1.43020	1.0043

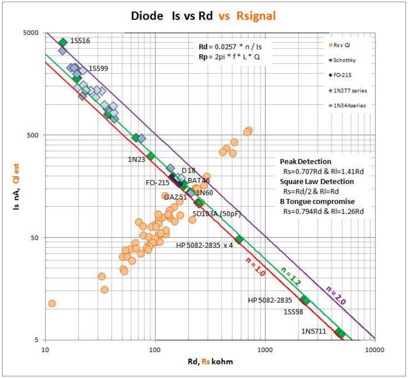
Averages: n = 2.01 Is = Tr nA Rd = 3.8E19 kOhm

The above spreadsheet is based on measurements of the diodes shown above. For the determination of Is and n, I chose at random two examples from my collection of various diodes, (all led's in this case) and measured with a modified methodology to that outlined by Ben Tongue and Mike Tuggle. In this case, with the radically different forward voltage drop of these diodes from Germanium or Schottky diodes, I have kept the Id values constant (about Id2=0.5uA and Id1=1.0uA) and varied Vd. I have included the actual room temperature in the calculations in order to eliminate this variable as a source of doubt or error. All the measured parameters as well as the determined results are presented in order to facilitate repeatability.

Why know the values of Is, n, and Rd of your diodes?

I have been attempting to get a handle on the main diode parameters Is, n, and Rd and how they impact the operation of a crystal radio. My method of choice is to use a graphical approach, plotting charts where variations between plotted parameters become apparent and easy to visualize. I have developed an interesting chart from the measurements above that compares many diodes with respect to Is, n, and Rd, and to understand how this fits into a matching situation with a fuzzy indication of typical tank impedance limits.

Ultimately one seeks a diode with an Rd that matches with and conjugate to the impedance it sees from the tank (+antenna). In the following discussions I will start with the cross-plot of Is to Rd and then proceed to a look at how received RF power plays a part in the selection and why low Is and n are desired.



The plot above shows the relationship between diode Is and Rd. Rd is calculated via the equation Rd = VT * n / Is where VT is the thermal voltage = k * T / q = 0.0257V at room temperature and Is and n are from the measured diodes. k is Boltzmann's constant = 1.38E-23 J/K and q is the electron charge = 1.609E-19 coulombs. On such a plot I can show lines of constant n, and plot the values for individual diodes for which I have spent considerable effort to determine the parameters Is, n and Rd (where Is is the diode saturation current, n is the ideality factor, and Rd is the diode resistance). On the plot it should be evident how changes in n or Is impact a diodes' Rd.