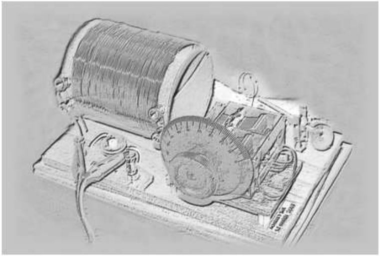
Crystal Radio Set Systems:

Design, Measurement, and Improvement

Volume I

A web book by Ben Tongue



First published: 10 Jul 1999; Revised: 01/06/10

NOTES:

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

PREFACE

The main purpose of these Articles is to show how Engineering Principles may be applied to the design of crystal radios. Measurement techniques and actual measurements are described. They relate to selectivity, sensitivity, inductor (coil) and capacitor Q (quality factor), impedance matching, the diode SPICE parameters saturation current and ideality factor, audio transformer characteristics, earphone and antenna to ground system parameters. The design of some crystal radios that embody these principles are shown, along with performance measurements. Some original technical concepts such as the linear-to-square-law crossover point of a diode detector, contra-wound inductors and the 'benny' are presented.

Please note: If any terms or concepts used here are unclear or obscure, please check out Article # 00 for possible explanations. If there still is a problem, e-mail me and I'll try to assist (Use the link below to the Front Page for my Email address).
Second note: The two dates following the Article titles are, respectively, the original publication date and the date of the last revision.

Original web location for book:
http://www.bentongue.com/xtalset/xtalset.html

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/ .

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.

TABLE OF CONTENTS:

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.

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 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.

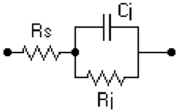
* The voltage readings are very sensitive to diode temperature. You can see this easily by grasping the diode body with thumb and forefinger and noting the change in the voltage reading when measuring V1 or V3. Don't take data until the readings stabilize. Saturation current is a strong function of junction temperature. For germanium and the usual (n-doped) Schottky diodes, a temperature increase of 10° Celsius results in a saturation current increase of about two times. A simple rule is: For each 1° 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.

ARTICLE 00

Practical design considerations, helpful definitions of terms and useful explanations of some concepts used in this Site

1. An explanation as to why some diodes that work well in a broadcast band crystal set 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.



Diode Schematic

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 RF tank. The tank usually has a small value tuning capacitor itself, so the overall tank circuit Q is reduced at high frequencies. This is the main reason why diodes having large values for Rs and CJ perform poorly at high frequencies.

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

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.

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 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 source, so we can say we have a 'no loss' situation. Now,

a conventional 3 1/2 digit DVM. It is also used when 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 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.

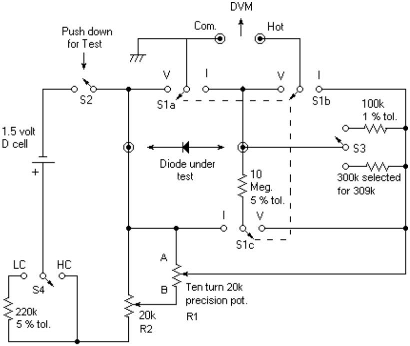


Fig. 1 - Schematic of Device for Measuring Diode Is and n.

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

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

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

ARTICLE 16

A Procedure for Measuring the Saturation Current and Ideality Factor of a Diode, along with Measurements on various diodes

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.

Warning: Don't use two diodes in series if you want the best weak signal sensitivity. The result of using two identical diodes in series is the emulation of an equivalent single diode having the same Is but an n of twice that of one original diode.

Experimental measurements on eleven different diodes used as detectors is shown in Article #27. Close correlation between these equations and actual measurements is demonstrated.

#15A Published: 04/10/2001; Revised: 12/07/2008

over the full range of currents and voltages encountered in crystal radio set operation, but varies with diode current in silicon p-n 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 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. Refer to the second schematic in Article #1. The output load will be considered to be a resistor connected across the "Audio Output" terminals; call it Rl. No audio transformer is involved in the first part of this discussion. 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 exist an appreciable power loss in the detector. The input and output resistances of the diode detector will approximately equal each other and approach Rd (diode axis-crossing resistance)= 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. This 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 well matched values. The input resistance of the diode detector decreased from the value obtained in the well matched low power situation. The output resistance increased. The reason for this change is that a new law now governs input and output resistance when a diode

as a function of the diode axis-crossing resistance and ideality factor (for weak signals and matched conditions).

Po={(Pi/n)^2}*(Rx/0.010554) (8)

Let us review the conditions that apply to equation (8) before discussing it. 1) The signal power to the detector diode is well below the LSLCP (weak-signal reception). 2) Detector source and load impedances are well matched to Rx, the axis-crossing resistance of the diode (see "Definition of Terms"). Equation (8) indicates that the maximization of output power, when receiving weak signals, requires using a diode of low n because output power is shown to be inversely proportional to the square of n. Also, with other factors being fixed, higher source and load impedances increase output power in direct proportion to their increase in value (remember they are both kept equal to Rx). A real-world problem occurs when one manipulates the RF input and audio output impedance transformations when attempting to optimize them to get greater weak-signal volume. Their internal power losses can change and this is not accounted for in equation (8). The RF input impedance transformation loss probably won't change a lot if the ratio of unloaded to loaded Q of the tank doesn't change, and will not be treated here. Audio transformer loss is worth investigating. An audio transformer usually has a Manufacturer-stated loss that applies when it is used between the Manufacturer-stated impedances. When used between other impedances (higher, in the usual case with crystal sets), transformer loss will most always be greater. Equation 8 shows that the transformer winding connected to the audio output side of the diode detector should present the highest impedance possible but, then that is done, transformer power loss usually increases. A compromise is needed.

reduced. At low input powers, the required input power for a specific output power approaches direct proportionality to the square root of n and/or Is, as shown in equation (5Li). The product n*Is can be considered to be a 'figure of merit' for diodes as weak signal detectors, provided input and output impedance matching exist and losses from passive components remain unchanged. The ideality factor (n) and saturation current (Is) of the diode are important parameters in determining ultimate very weak signal sensitivity. If all other diode parameters are kept the same, the weak signal input and output resistances of a diode detector are directly proportional to n and inversely proportional to Is. Assume a diode with a value of n equal to oldn is replaced with an identical diode, except that it has an n of newn, and the input and output impedances are re-matched (the new impedances are doubled). The result will be a detector insertion power loss change of: 10*log(oldn/newn) dB. That is, a doubling of n will result in a 3 dB increase in insertion power loss, assuming the input power is kept the same and input and output impedances are re-matched. The result is a 3 dB reduction of output power (volume). A similar effect occurs if Is of the diode is increased except that this change reduces the impedances that must be re-matched instead of increasing them.

Interesting note: A simple manipulation of Equations #0, 2 and 4 shows that, at the LSLCP, the RMS value of the AC signal at the input to the diode is: (0.08895*n) Volts, and it is independent of Is.

One final equation: One can combine two equations: Rx=0.0256789*n/Is, as given in the "Definition of Terms" at the beginning of this Article and equation (5Lo) to come out with a very useful result relative to actual crystal set design. For a given input power, equation (8) gives the output power

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 resistance. Further, since the detector is now a peak detector, the DC output voltage approaches 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 RF 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). Note: It is assumed here that the peak reverse voltage applied to the diode when the signal is strong does not approach its peak reverse breakdown voltage rating.

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 (Rd=axis-crossing resistance of the diode). 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 the signal level. This situation can be attained in a real-world crystal set using an audio transformer if one connects a parallel RC (a "benny") in series with the transformer primary. The resistor plus the DC resistance of the transformer primary is adjusted to equal the average AC resistance looking into the transformer primary. This combination approximates the effect of a straight resistive load, as used in the discussion above. To determine the average AC input resistance of the audio transformer (when loaded by phones), multiply its step-up impedance ratio by the average AC impedance of the phones. See Article #2 for a method of determining the average AC resistance of phones.

In the practical case of a real-world crystal set not using a "benny", the load the diode sees has a DC resistance component lower than its AC impedance. This unbalanced condition is worse when a transformer is used, as compared with phones only. In this case, when going from reception of a weak to a strong signal, bad things start happening to selectivity: As the RF signal strength starts increasing, the rectified DC current in the diode starts increasing faster than it does in the condition using a "benny". This causes the diode input and output resistances to fall. The result is a loaded down tank having reduced selectivity. The use a "benny" prevents this, among other things from happening.

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

From equation (5r), at low output power power levels, the input power required to produce a given output approaches:

Pi=sqrt(0.41104*n*Is*Po) (5Li)

Prearranging terms of equation 5Li:

Po=(Pi^2)/(0.41104*n*Is) (5Lo)

An equation that fits the detector Insertion Power Loss Ratio (DIPLR) is obtained by dividing equation (5) by Pi:

DIPLR=[sqrt(1+0.102716*n*Is/Pi)-sqrt(0.102716*n*Is/Pi)]^2 (6)

DIPLR={sqrt[1+Plsc(o)/Pi]-sqrt[Plsc(o)/Pi]}^2 Normalized to Plsc(o) by dividing right side of eq. (5n) by Pi (6n)

The Insertion power loss (DIPLR) at which a diode detector is operating can be determined by noting the DC rectified output voltage:

Adjust the DC component of the diode load to equal its axis-crossing resistance (0.0256789*n/Is ohms).** Measure the DC voltage V2 developed across the DC load. Some simple manipulation of equation (3) above results in equation (7).

DIPLR=V2/(V2+0.1027156*n) (7)

** See articles #16 and #27 for info on determining the Is and n of diodes as well as measurements on some diodes.

Equation (5r) seems important. It shows that the input power required for a specific output power is reduced if n and/or Is is stated earlier. Curve fitting suggests this relationship: Po/Pi=(I2/(I2+4*Is)=DIPLR (3)

Interesting note: Since, at the LSLCP, I2=2*Is (eq. 1), Plsc(i)=Plsc(o)*3. (4)

Calculation of the diode detector input power when it is operating at its LSLCP:

Substituting the value of Plsc(o) from equation 2 into equation 4 results in: Plsc(i)=0.308148*Is*n for the input power at the LSLCP (4a)

Various relationships between input power, output power and diode parameters:

A lot of mathematical manipulation of the relations given above results an equation that fits the simulation data quite well over the whole range of the graph in Fig. 2.

Po=[sqrt(0.102716*n*Is+Pi)-sqrt(0.102716*n*Is)]^2 Watts. (5)

Po={sqrt[Plsc(o)+Pi]-sqrt[Plsc(o)]}^2 Watts. Normalized to Plsc(o). (5n)

A rearrangement of the terms in equation (5) yields:

Pi=Po+sqrt(0.41104*n*Is*Po) Watts. (5r)

Pi=Po+2*sqrt(Plsc(o)*Po) Watts. Equation (5r) normalized to Plsc(o), by using equation (2). (5rn)

impedance-matched! There is little benefit when strong 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 regarding 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.

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

Calculation of diode detector output power when it is operating at its LSLCP:

Differentiating the Shockley diode equation with respect to the diode junction voltage yields:

Diode junction resistance=Rx=0.0256789*n/Is ohms. (0)

At the LSLCP, I2=2*Is. (From the paragraph above) (1)

Some obvious relations: Output power=Po=(I2^2)*R2 Watts. The output load R2 has been specified as equal to Rx, Rx is defined in equation (0). I2 is 2 times Is at the LSLCP (See equation 1). Substituting into the equation for Output power Po, we get, for the output power at the LSLCP:

Plsc(o)=0.102716*Is*n. (2)

Calculation of the diode detector DC output voltage [V2lsc(o)] when it is operating at its LSLCP:

Some simple manipulation of equations (0) and (2) results in the relation: V2lsc(o)=0.051*n Volts (2a)

Calculation of the ratio of output power to input power (DIPLR) at any power level:

A proper relation between Po, Pi and I2 obviously requires that I2 approach zero as Po/Pi approaches zero, that Po/Pi approach proportionality to I2 as I2 becomes low (the square law relation) and that Po approach Pi as I2 becomes very high. Also, at an output power of Plsc(o), I2 must equal 2*Is, as

S11 approaches about -9.5 dB at very high input power levels. The input impedance match conditions are reversed in Case B. Rx is still 696009 ohms but R1 is set to 492153 and R2 to 984305 ohms. Now, at low input power levels, the input and output are somewhat mismatched (S11=-15.3 dB), but at high signal power levels, a very good impedance match is approached. The bottom row shows these two conditions.

Unpublished information from Xavier Le Polozec indicates that the optimum practical compromise values for R1 and R2, for input powers well below the LSLCP to well above it is: R1=Rx and R2=Rx*sqrt2.

The diode detector equations:

The following equations are developed for the Class A termination condition of R1=R2=Rx. They give insertion power loss values to within a fraction of a dB of those provided by SPICE simulation for both Cases A and B. The output current, I2, is different for Class A and B.

Observation of a curve of output vs. input power (in dB), from SPICE simulation of a Class A terminated detector reveals a slope of 1.5 at an input power value of -78.91 dBW, for the particular diode used. This is the LSLCP for input power. Another observation is that the rectified diode current I2 at the LSLCP point appears to be very closely two times the Is of the diode. This two times figure appears to be apply to all diodes. Note: The Plsc(i) of -78.91 dBW occurs when the detector parameters are R1=R2=Rx and the diode parameters are: Is=38 nA, n=1.03 and temperature is 25° C. In general, the 'two times' figure does not hold for termination conditions other than R1=R2=Rx.

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 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 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.

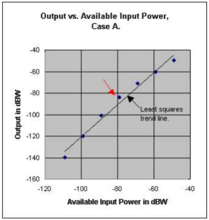
There are power losses at each of the three steps and our job is to minimize them in order to get as much of the sideband RF power passing through the vicinity of the antenna (capture area) changed into audio power for our ears. We want all of the "available power" at the antenna-ground system to be absorbed into the crystal radio set then passed on through it to our headphones as sound. However, some of it will be

Note the following in Fig. 2: At input power levels well above the LSLCP, the relationship between input and output power (see the data points, not the Least Squares Line) approaches linearity. That is, the output changes about one dB for every one dB change in input. At input power levels well below the LSLCP, the relationship between input and output power (see the data points, not the Least Squares Line) approaches a square law relationship. That is, the output power changes about two dB for every one dB change in input power. This has bad implications for weak signal reception. If a weak signal fades, the detected signal will drop twice as many dB as the reduction in input signal strength. For best weak signal sensitivity, one should push the LSLCP to as low a power level as possible. This moves weak signals closer to the LSLCP and linear operation (less detector power loss), and thus increases volume. Lowering of the LSLCP power is associated with using a diode having low values for n and Is, and impedance matching the antenna-ground system impedance and headphone impedance to the now higher values of detector input and output impedances. For good results, one must make sure the impedance transforming means does not introduce other losses. A high Q tank inductor, tuning capacitor, and low loss audio transformer are important. It may be difficult to achieve the required greater impedance transformations in a low loss manner.

Example: Assume that Is=38 nA and n=1.03, as was used in the graphs and chart above. Rx becomes approximately 696009 ohms. R1 and R2 are each fixed at 696009 ohms for all simulation points in Case A. This establishes a very good input and output impedance match at low signal levels and a moderate match at higher levels. The input return loss (S11) is better than -14 dB at signal levels up to 12 dB above Plsc(i).

mismatch exists at both the input and output ports. In Case B, conversely, the detector input and output ports are both are moderately impedance mismatched when the signal power is several times lower than that at the LSLCP. At signal power inputs several times greater than that at the LSLCP, both input and output ports approach an impedance matched condition.

Simulated diode detector output power and insertion loss vs input power, Case A.
This diode has a low saturation current, compared to that used in the average
crystal set, Is=38 nA, n=1.03 The LSLCP is shown by the red arrow.
Graph of output vs input power Graph of power loss vs input power



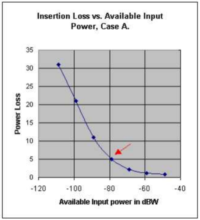


Fig. 2 - A SPICE simulation of the relation between output and available input power.

Fig. 3 - Data from a SPICE simulation showing detector insertion power loss vs. input power.

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.

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 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.

6A. A Short tutorial on some aspects of audio transformer utilization in crystal sets. See Section 3 of Article 14 here

7. A caution to observe when cutting the leads of a glass Agilent 5082-2835 Schottky diode (as well as some other types of glass diodes): 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 when the cut happens. This can damage the electrical performance of the diode. The 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. This reduces the possibility of heat damage during soldering. The wire must also have close to the same temperature coefficient of expansion as the glass to enable a gas-tight seal to be obtained.

This power loss figure changes by less than 0.1 dB between Cases A and B.

Most crystal radio sets can deliver a readable signal at an input of Plsc(i) Watts. It would obviously be desirable to lower the input power at which the LSLCP occurs so that more of the weak signals would be closer to the linear mode of operation, experience less insertion power loss and therefore be louder.

Example SPICE simulation of a diode detector at 25° C.: Figs. 2 and 3 show power relations at various power levels. The LSLCP is shown by a red arrow. In the SPICE simulations for these graphs, the source and load resistances, R1 and R2, are equal to Rx (Case A). The DIPL values for Case B (Fig.3) are within 0.4 dB of those of Case A. The main difference is the lower DIPL values in Case B at high input powers. For instance, the insertion power loss at an input of -48.912 dBW is 0.76 dB for Case A and 0.30 for Case B. The loss figures from the equations that follow are quite close to those that occur in a SPICE simulation of both Case A and Case B.

At input power levels several times or more below the LSLCP, the impedances of the input and output ports of the detector both approach Rx for both Cases, A and B. At input power levels several times or more above the LSLCP, the detector approaches operation as a peak detector having a low insertion power loss. In this condition the input RF resistance of the detector approaches half the output load resistance and the output resistance of the detector approaches twice the RF source resistance. Summary: In Case A, the detector input and output ports both approach an impedance matched condition when the signal power is several times lower than that at the LSLCP. At signal power inputs several times greater than that at the LSLCP, a moderate impedance considered. In Case A, R1=R2=Rx=0.0256789*n/Is. In Case B, R1=Rx/sqrt2 and R2=Rx*sqrt2.

1. The low power region: Here, the relation between output power and input power approaches 'square-law'. That is, for every one dB change in input power there is about a two dB change in output power. The detector input and output resistances approximate Rx.

2. The high power region: Here, the relation between output power and input power approaches 'linear'. That is, for every one dB change in input power there is about a one dB change in output power. The detector input and output resistances are no longer equal. The detector input resistance is equal to about half of R2. The detector output resistance is about twice R1.

3. The point where the two areas overlap equally: This is the 'linear-to-square-law crossover point' (LSLCP). At this point there is a 10*log(sqrt2) dB change in output power for every 1.0 dB change in input power (slope of about 1.5). If R1 and R2 are both equal to Rx, in Case A the detector input resistance is about 12% less than Rx and its output resistance is about 12 % greater than Rx.

Transition from the linear to the square law region: All good diode detectors, at high input power levels, if well impedance matched at input and output, have a low insertion power loss (a fraction of a dB). If the input power is reduced, at first the output will drop approximately dB for dB in step with the input. If the input is further reduced, the output will start to drop faster (in dB). This can be thought of as the onset of noticeable 'detector insertion power loss'. The insertion power loss at the LSLCP, in two SPICE simulations (see Cases A and B below), is about 5 dB. Put another way, at the LSLCP the output power is about 0.3 times the available input power.

The appropriate type of wire is called 'dumet'. One can read about it at http://www.dumet.net .

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 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 out the harmonic and sum components as well as the two original signals from the output in a non-dissipative manner, 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 an AC signal power source of P Watts 'available power' is impedance matched to the input AC impedance of the diode detector, the DC output power can closely approach the 'available power, P watts', 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 CW AC power of frequency 'f' Hz into output power of frequency zero Hz (DC). This is called the detected DC output. If the input power is large enough, the power loss in the black box can approach zero. Reduced input power levels result in an increased insertion power loss in the AC to DC power conversion of the "black box).

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 #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

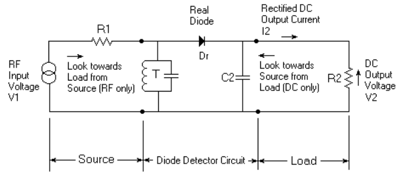


Fig. 1

Assumptions used in the following discussion:

* The Q and L/C ratio of tuned circuit T are assumed to be high and low enough, respectively, so that the 'stored energy effect' of T prevents any appreciable clipping of the positive voltage wave form peak by diode D1.
* The value of C2 is assumed to be high enough so that a negligible amount of RF voltage appears across it.
* The diode parameters Is and n are known from measurement or a Data Sheet. A simplified method of estimating Is is given in Section 2, Article #4, but the parameter n has to be estimated. A method for measuring both Is and n is given in Article #16. The effect of the series parasitic resistance of the diode is assumed to be negligible - as it is at low signal levels for most all detector diodes. Diode back leakage current from either 'parasitic leakage' or operation with voltage swings reaching into the 'reverse breakdown current' region is assumed to be negligible. The diode temperature will be assumed to be 25 degrees C.

Approach: The RF signal input power range is divided into two regions and one point; impedance and power relationships are determined. Refer to Figs. 2 and 3. Two Cases will be

Plsc(i) Input power at the linear-to-square-law crossover point
Plsc(o) Output power at the linear-to-square-law crossover point
Is Saturation current of the diode
n Ideality factor of the detector
DIPL Detector insertion power loss in dB
DIPLR Detector insertion power loss ratio (ratio of output to input power)
DIR Detector input resistance (AC)
Pi Available input power
Po Output power
sqrt Take the square root of the expression following
Kt Temperature in degrees Kelvin
C. Temperature in degrees Celsius
Ri Detector input resistance
Ro Detector output resistance
R1 Source resistance
R2 Load resistance
I2 Rectified current
Rx Slope of voltage/current curve of a diode at the origin. Rx=0.0256789*n/Is, at 25° C.
S11 A measure of input impedance match. S11=20*log|[(|Ri-R1)/(Ri+R1)]|. S11 is always a negative number, and the greater its absolute value, the better the impedance match.
V2lsc(o) Detector DC output voltage when it is operating at its LSLCP
SPICE A computer circuit simulation program. ICAP/4 from Intusoft was used in all simulations.

The diode detector circuit to which we will refer is shown in Fig. 1.
Schematic of diode detector circuit

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 (making it sound brighter or tinny) and reduce the bass response, compared to the response when the AC source resistance matches the effective impedance of the elements. Conversely, driving the headphone elements from a low resistance source tends to roll off the treble (making it sound dull) and relatively speaking, increase 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.

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 Figs. 2 & 3 in Article #15A. See the last paragraph of Article #17 for a way to determine if a diode detector is operating below or above its LSLCP.

* 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.

* 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 series-connected capacitor.

* Refer to Articles #2, #3,#5 and #14 for more info. For a convenient way to vary the audio impedance driving the phones, consider the 'BT-Ulti-Match' described in Part 4 of Article #5.

ARTICLE 15A

Quantitative insights into Diode Detector Operation derived from Simulation in SPICE, and some Interesting new Equations relating diode parameters to weak signal sensitivity

Quick Summary: Several new equations are presented showing various relations between diode detector rectified current, input AC and output DC power, insertion power loss and the 'Linear-to-Square-Law Crossover Point' (LSLCP). The LSLCP is an operating point where the diode detector is operating half way between its linear and square law modes. It can also be thought of as a dividing point between weak and strong signal reception. Bear in mind that the LSLCP is a point (see Fig. 3) on a graph of output DC power vs input RF power of a diode detector system. It is not a point on a graph of DC current Vs voltage of a diode. Article #27 shows actual measurements on a crystal radio set using eleven different diodes, that tends to experimentally back up the validity of equation #5 and those following it.

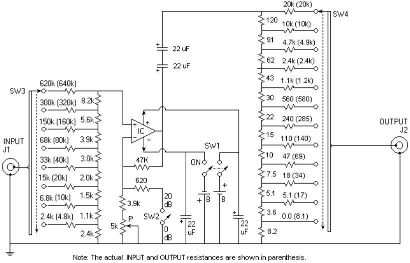
This Article, #15A, used to be Part 1 of the old Article #17.

Definitions of terms to be used:

Class A Impedance matching condition in which R1=R2=Rx

Class B Impedance matching condition in which R2=2*R1 and sqrt(R1*R2)=Rx

LSLCP A point on the curve of output power vs input power of a diode detector where it operates half way between its linear and square law mode



Note: The actual INPUT and OUTPUT resistances are shown in parenthesis.

Fig. 4 - Schematic of a 'Unilateral Ideal Transformer' Simulator, ver. B.

Schematic diargam

#14 Published: 01/05/01; Last revision: 01/17/10

Note: Bullet points 2 and 3 will also change tone quality when receiving strong signals.

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

From my early experience in the manufacturing 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 in value more, over time, than the low value ones.

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 A-B 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 and were probably made by them.

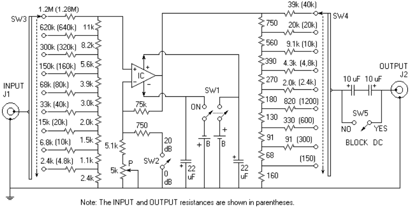
A fact of interest that some may not know is this: The AC resistance of carbon composition resistors, and film resistors, to a much lesser degree, decrease with increasing frequency (the Boella Effect). This effect is strongest in high value

If no audio generator is available, connect the output of the crystal radio set diode detector to J1 (no audio transformer to be used), and a headphone set of about 10k ohm impedance (2k ohm DC resistance) to J2. Tune in a station and adjust potentiometer P so that the volume is the same as when the detector output feeds the headphones directly. This setting does not have to be changed in the future. Note: Connect the output of the crystal radio set detector to the UITS with as short a length of cable as possible in order to minimize added shunt capacity. If the tone quality of the signal changes from one resistance setting of SW3 to another, the shunt capacity in the detector output circuit is too high. This can be caused by using a diode RF bypass capacitor or an interconnecting cable of too high a shunt capacitance for the resistance setting of SW3 being used. I use an eighteen inch length of RG-59 type coax for my cable. It has a capacitance of about 20 pF per foot.

The performance of magnetic diaphragm type headphones can be affected by the DC current passing through them when no coupling transformer is used. SW5 is provided for those who choose to block the DC.

6. Schematic for version B (Added 05/25/2003). The differences between this version and that shown in Fig. 3 are:

1. The output resistance range is changed from 40k-150 ohms to 20k-8 ohms. This allows the use of the UITS with typical dynamic headphones.
2. The DC blocking is made fixed (SW5 is eliminated).
3. The schematic shows only provision for input impedances up to 640k. The extra switch position for 1.28M shown in Fig. 3 may be added if desired.



Note: The INPUT and OUTPUT resistances are shown in parentheses.

Fig.3 - Schematic of an 'Ideal Transformer' Simulator.

Schematic of original UITS

Some component specifics:

* B: 9 Volt batteries.
* IC: JFET input op-amp such as one section of an LF353, TL081 or M34002. Basically, it should have a JFET input and a gain-bandwidth product of 3 MHz or more.
* The 22 uF caps, electrolytic or tantalum, should have a voltage rating between 10 to 25 Volts.
* The resistor values shown in the schematic are those in the standard 5% series of values. The use of resistors that difer by +/- 10% from the values shown should not have an appreciable impact on performance of this unit.

5. Setup.

Calibration is simple. With SW2 in its 0 dB position and SW3 and SW4 at their 10k Ohm settings, set potentiometer P for zero gain. To do this, load J2 with a 10k Ohm resistor and feed a 1 kHz signal from an audio generator into J1. Adjust P so that the output voltage at J2 equals the input voltage at J1.

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

An advantage of the old carbon comp resistors over the newer carbon-film and metal-film resistors is the fact that they can withstand much greater momentary power overloads. After being out of the carbon comp resistor business for many years, Stackpole re-entered it a couple of years ago for this very reason. For the same power rating, they usually have a higher voltage rating.

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.

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.

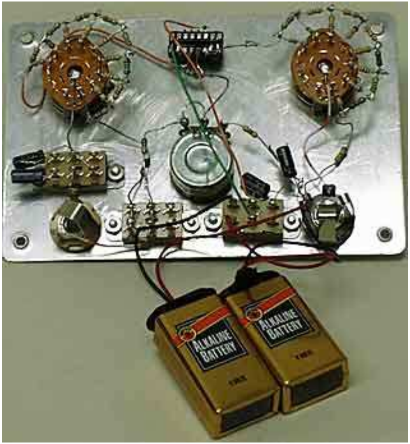


Fig. 2 Underside view of UITS

and secondary source and load resistances about equal to the values determined with the UITS. A transformer that has factory specified impedance levels as much as four times lower than desired, but with the correct transformation ratio, and a frequency response range much wider than 0.3-3.3 kHz will probably work well.

Note. The parallel RC (a 'benny') (see Article #5), needed in series with the primary of a real world transformer, is not needed with the UITS because its input resistance is the same for DC as for AC.

Fig 1. Front view of UITS

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. 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 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

How should one proceed in determining the specifications for a transformer that will provide optimum performance in the crystal radio set? One may not know the audio source resistance of the diode detector, or even the average impedance of the headphones load. The UITS can be used to find these two values. It also has a 20 dB gain switch option that can be used to enable reception of very weak signals as well as a switch to block DC from the phones, if desired. There are two operating adjustments. One sets the input resistance Ri, the other the output resistance Ro. These two settings don't interact. The equivalent real-world transformer turns ratio is the square root of the ratio of the two resistance settings. Here are some ways that the UITS can be used:

* Compare the performance of a candidate transformer to that of an ideal transformer to see how much signal is lost in the candidate. There is no point in looking for a better transformer if the difference between the two is small.
* Use it to find the impedance transformation ratio that would be optimum for the crystal radio set/headphone combination being used.
* Use it in place of an actual transformer.
* Enhance reception of very weak signals.
* See bullets in the "What's it good for?" section, above.

To use the UITS, connect it between the detector output and headphones. Insure that the diode has an appropriate RF bypass capacitor. Set the amplification to 0 dB. Adjust each rotary switch independently for the loudest volume. Calculate the impedance transformation ratio from the settings of S3 and S4. A transformer specified with this ratio is optimum for the detector and headphone impedances being used, all other things being equal. Its specifications should include primary with the primary and secondary windings.

It is important to understand that the impedance numbers a transformer manufacturer specifies for various terminals are the source and load resistances they used to specify the performance of the transformer. In most instances when one uses a transformer originally designed for use between tubes (plate-to-grid) or to match a low impedance line to a plate or grid in a high-performance crystal set application, the following statement applies: Optimal results usually occur when the actual crystal set source and load resistances are higher than the values for which the transformer was designed, but are of the desired ratio. For instance, if a Stancor A-27 transformer is connected with terminals 8 and 9 joined together and 3 and 4 joined together (diode input to pin 7, diode return to pin 10), it is specified by the manufacturer for use between a 100k source and 600 ohm load connected to pins 1 and 6. Using it between approximately 200k and 1.2k ohms will result with a very slight increase of midband transformer power loss, but a greater offsetting increase of detector sensitivity from the higher detector load of 200k. This assumes a diode having a relatively low saturation current such as a Schottky, FO-215 or other high axis-crossing-resistance diode is used. The audio bandwidth will also be reduced but probably by a non-perceivable amount. Some illustrations of these effects can be viewed in Article #5, Tables 5, 7 and 8. To go further in this direction to present the diode with an even higher transformed resistance from a 1200 ohm load, connect the load to terminals 1 and 5 instead of 1 and 6. Volume may diminish from increased transformer loss and reduced audio bandwidth.

4. The Unilateral 'Ideal Transformer' Simulator.

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

13. Comments on audio distortion in diode detectors.

There seems to be four main causes of audio distortion in diode detectors.

* Square-law detection: Second-order audio distortion occurs when receiving weak signals. This type of distortion is usually not irritating to the ear.

* Distortion when the signal is strong: This type of distortion occurs when the DC load on the diode is very different from the average value of its AC load. This is usually the case when an audio transformer is used without load resistance correction. The remedy is to use a ”benny” to equalize the DC resistance of the diode load to equal that of its average AC load. See second paragraph of Part 2 in Article #1. This type of distortion also occurs when using a piezo earphone without an appropriate shunt resistor.

* Distortion from using a diode having too low a saturation current (high axis-crossing resistance). This usually occurs when receiving weak signals: The remedy is to modify the crystal set to raise its audio load impedance (change the audio transformer) to achieve better impedance matching to the diode. Another approach is to forward bias the diode to lower its axis-crossing resistance (raise its effective saturation current) to better impedance match the existing audio load. See Article #9. Another solution is to use a different diode having an axis-crossing resistance that more closely impedance matches its source and load. See Table #2, second column for the axis-crossing resistances of some diodes.

* Distortion when using a diode having a low reverse breakdown voltage and having a sharp knee at breakdown (usually a microwave diode). The peak reverse voltage on a diode can approach twice the detected DC voltage. When a signal is strong enough to cause this to happen on the peaks of modulation, a large reverse current can be drawn (reverse current is normally low) that clips the peaks of the detected envelope causing audio distortion and a weaker detected signal. A remedy is to change to another type of diode or reduce the strength of received strong signals. See attenuators SW1 and SW2 in Fig. 5, in Article #22.

Some measurements of saturation current and ideality factor values for about 15 diodes be found in Articles #16 and 27.

A suggested definition for “strong” and “weak” signals: A diode detector may be considered entering the "strong signal" mode of operation if it is operating at a signal power level about 10 dB or more above its LSLCP (Linear to square law crossover point). One can estimate that the diode detector is entering the "weak signal" mode of operation if it is operating at a signal power level about 10 dB or more below its LSLCP.

To illustrate this point we will consider a transformer designed to transform a 10,000 Ohm source to a 90,000 Ohm load. This could be an AES PT-156, Stancor A-53C or similar transformer originally designed to couple the output of a first (tube) audio stage to push-pull grids. If the designer did a good job, this transformer will have the lowest possible loss consistent with its specified frequency range, power handling capability and cost goals. If it were to be driven from a 40,000 Ohm source and loaded with a 360,000 Ohm load (still a 1:9 impedance ratio), its center-band power insertion loss will be increased and the low frequency end of the band will be rolled off. The reason for the increase of center-band loss is that the shunt resistance caused by losses in the iron core load down the now higher source resistance (40,000 Ohms) thus increasing loss. The shunt inductive reactance of the primary winding, at the low end of the band loads down the now higher source resistance (40,000 Ohms) more than before, thus increasing the roll-off at the low frequency end of the audio band. The high end of the audio band will also probably be rolled off because the reactance of the shunt capacitance of the primary winding will cause more loss when being driven by a 40,000 Ohm source than one of 10,000 Ohms. On the other hand, if the transformer was driven from a 2,500 Ohm source and fed a 22,500 Ohm load, center-band power insertion loss again still be increased. The reason is the ratio of the source resistance to the series resistance of the primary winding is not as high as when the source was 10,000 Ohms. More of the input power will be dissipated in this series resistance and less transferred to the secondary. A similar loss effect from the winding resistance occurs in the secondary. The low frequency end of the band will reach to lower frequencies than before, but the high end may get some roll-off due to leakage inductance in the primary and secondary windings. One can think of this effect by visualizing a parasitic inductor in series from J2 back to J1. The 'unilateral' in the name comes from this property. See Fig. 3. A real world transformer is bilateral. That is, it can pass a signal in either direction.

* A good transformer has very little loss. The UITS can be set to have no power loss (or gain), no matter what the effective turns ratio setting is. The effective turns ratio is controlled by the settings of SW3 and SW4. A real world transformer has a turns ratio of, say 'n'. This gives it an impedance transformation ratio of n^2. That is, a resistor of value R, connected to one winding will be reflected as a value R*(n^2) or R/(n^2) at the other. 'n' is a fixed parameter of the real world transformer unless it has taps, then several various values of 'n' can be obtained. The UITS can be adjusted with SW3 and SW4 to a very wide range of transformation ratios. It has the advantage of independent control of input and output resistance by means of switches, with no power loss for any combination of input and output resistance.

3. A Short tutorial on some aspects of audio transformer utilization in crystal sets.

One of the issues one encounters when designing a high performance crystal radio set is determining the optimum parameters for the detector-to-headphone audio coupling transformer. Its impedance transformation ratio is the main factor to be considered, though the inherent loss and reactance parameters are also important. Another factor is the primary and secondary impedance levels for which the transformer was designed, compared with the levels to be used in a crystal radio set application.

Consider the performance of two transformers having the same transformation ratio, but originally designed to operate at different impedance levels. They will not perform the same.

A discussion of LSLCP may be found in Section 1, Article #15. A discussion of diode operation at its LSLCP may be found at the end of Article #17.

#0 Published: 04/25/00; Last revision: 01/01/11

ARTICLE 01

A New Way to look at Crystal Radio Design. Get Greater Sensitivity to very Weak Signals, and Greater Volume, less Audio Distortion and Improved Selectivity on Strong Signals

Quick introduction:

Greater sensitivity to very weak signals can be attained by lowering the RF signal power level (linear-to-square law point, or LSC point) at which the detector changes from the linear to the square-law mode of operation (See Article #10, Figs. 3 & 4 and part #3 for an explanation of the LSC point). This is accomplished by connecting the highest impedance point of the RF tuned circuit to a diode having the proper Saturation Current (See Article #15A). The output resistance of the detector should be impedance matched to the headphones, usually by a low-loss audio transformer, for maximum sensitivity. Greater volume, less audio distortion and improved selectivity can be attained on strong signals by properly impedance matching the RF source resistance to the RF input resistance of the detector and also matching the output resistance of the detector to the effective impedance of the headphones. The DC and audio AC loads on the detector should also be made equal. This analysis does not involve the analysis of diode instantaneous voltage and current wave-forms, input voltage, output voltage, diode turn-on voltage or tuned circuit peak-clipping. This analysis does consider the detector to be a black box having a linear input RF resistance and a linear output resistance, is driven from an AC power source and delivers power to an output load. These resistances are independent of input signal power at low power levels (somewhat below the LSC point) and depend only upon the

* The sensitivity of various headphones may be compared without the problem of needing an optimum audio transformer for each. Just adjust SW4 for maximum volume on each headphone and read the approximate optimum source resistance from the calibration.

* One can determine, in a particular crystal radio set, how closely a particular real-world transformer emulates an ideal one.

* One can easily demonstrate how the frequency response (tone quality) of a particular headphone changes as a function of the source resistance driving it by changing the setting of SW4.

* One can also find out out if one's real-world audio transformer alters tone quality. This can happen if its shunt inductance is too low or if its distributed winding capacitance is too high.

* The average audio impedance of headphones can be determined. For more info on this, see Articles #2 and 3.

* An added feature of the device as implemented is the capability of adding a 20 dB boost to the audio signal (this is where the plus... comes from). This feature does not affect the input and output resistances. It can be used to just add volume to weak signals, or as an aid in centering tuning on a very weak signal.

* In normal operation (20dB boost turned off), the UITS is calibrated to provide no power gain of loss. It has a flat frequency response +/- 0.3 dB over the audio band of DC - 3.3 kHz.

2. What is it?

* The UITS, unlike a real world transformer, can pass a signal from the input port (J1) to the output port (J2), but not

1. What's it good for?

Consider a crystal radio set that uses an audio transformer to drive headphones. One can determine what its performance would be if the transformer had no loss and provided an optimum impedance match between the output resistance of the diode detector and the headphone load.

One can determine if the optimum diode load resistance changes as a function of signal level by adjusting SW3 for the loudest volume on a weak signal and then readjusting it for a strong one.

* One can determine the optimum turns ratio for a real-world transformer. To do this, set SW3 and SW4 for maximum volume. Calculate the output-to-input winding turns ratio as the square root of the ratio of the port resistances of SW4 to SW3 (the numbers in parentheses in Fig. 3).
* One can determine if the optimum diode load resistance changes from one end of the BC band to the other by adjusting SW3. It usually does change, when receiving strong signals.
* Some of the mystery can be taken out of evaluating diodes. A diode will exhibit its best weak-signal sensitivity when the RF source resistance driving it, and the audio load resistance are set to the optimum values for that diode. When comparing various diodes in a crystal radio set that is using a Unilateral 'Ideal Transformer' Simulator (UITS), the optimum audio load resistance required for that diode can be easily dialed up just by setting SW3 for the loudest volume. The diode is then not penalized for being used in a poor impedance environment (for that diode).

characteristics of the diode. At high input power levels (somewhat above the LSC point), the input resistance is still linear and depends primarily on the output load resistance. The output resistance depends primarily on the source resistance.

1. THEORY

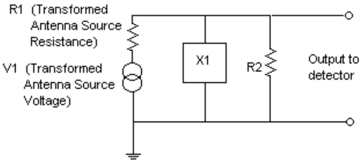
A crystal radio may be thought of as the cascaded connection of several basic components.

* Antenna-ground system: Signal source
* RF tuned circuit: Provides selectivity and impedance matching between the resistance of the antenna-ground circuit and the RF input resistance of the diode detector. This tuned circuit has some power loss.
* Diode detector: Characterized as a black box that accepts RF input power and converts it to DC output power. It has an RF input resistance, an audio output resistance and a power insertion loss (dB). These three characteristics are interrelated with the RF Input power, RF source resistance driving the detector, audio load resistance and the parameters of the diode used.
* Output transformer: To impedance transform the effective headphone impedance to that required by the diode.
* Audio load: Headphones, what else?

We will consider these components one at a time. See Part 1 of Article #10 for an overall view of the way we will be looking at diode detector operation.

The Antenna and RF Tuned Circuit will be combined into three components. V1 and R1 represent the antenna induced voltage and resistance, impedance transformed by the tuned circuits and antenna reactance to the series-connected values seen by the diode detector. X1 represents the reactance of the tuned circuit(s) seen at its output terminals. Its impedance is considered to be substantially zero at harmonics of the frequency to which it is tuned. Its impedance is also substantially zero at DC and at Audio frequencies. R2 represents all the losses in the tuned circuits at resonance, as seen by the diode. This is not the conventional way of viewing the signal source for a detector.

Schematic of Antenna and Tuned Circuit Load.



The Detector will be represented as follows: The LC tank assures that the input is effectively shorted to ground at DC and at audio frequencies as well as all RF frequencies except that to which it is tuned. The output is effectively shorted to ground at RF by C1.

Schematic of Detector with Source and Load Impedances.

ARTICLE 14

A Zero Loss, Unilateral 'Ideal' Audio Transformer Simulator, plus... This device makes is very easy to determine the optimum audio transformer source and load resistances for any crystal set diode/headphone combination. No test equipment necessary

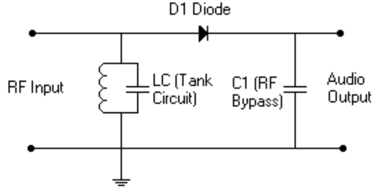
Quick Summary: This device works as an audio transformer when connected between the output of a diode detector and headphones, but with several differences. (1) No insertion power loss. (2) Input and output resistances can be independently varied over a wide range by selector switches. This provides for the simulation of a wide range of "transformer turns ratios".

The main purpose of this device is to enable oneself, by twisting two dials, to find out the optimum audio impedance transformation needed in a 'real world transformer', while experimentally trying different diodes or headphones in a crystal radio set. The effect the transformer has on selectivity and volume may be evaluated. Another purpose is to enable one to check how closely the performance of one's audio transformer conforms to that of an ideal one, both having the same input and output impedances. It also has a switchable 20 dB amplifier to enable better reading of very weak signals.

The first version of this device, shown in Figs. 1, 2 and 3 is designed for driving typical sound-powered balanced-armature, magnetic diaphragm or piezo electric earphones. The schematic for Version B, shown in Fig.4 is designed for feeding a wider range of loads, down to an impedance of 8 ohms. This unit can match the impedance of the earphones mentioned above as well as that of typical dynamic earphones.

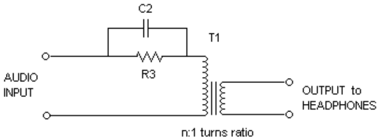
weaken and air gaps may get changed. All affect the sensitivity.

#13 Published: 08/30/00; Last revision: 10/28/2004



The Output Transformer circuit will be represented as shown below. The purpose of R3 and C2 will be covered later.

Schematic of Output transformer with RC series Element.



We start out with the assumption of no losses in the tuned circuits. This condition makes R2 equal to infinity, not a practical assumption of course, but it will simplify what follows. The input circuit then reduces to a simple series connection of the parallel tuned circuit, impedance transformed antenna voltage, and a series resistance. This resistance includes the effects of radiation, antenna, lead-in and ground circuit resistance. A simple transformation enables us to eliminate R2 entirely by combining its effects into a changed value for R1 and a new value for V1. The new value for V1 is: V1new = V1old*(R2old/(R1old + R2old)). The new value for R1 is: R1new = (R1old*R2old)/(R1old + R2old). With this transformation the new value for R2 is infinity, so it can be eliminated from the circuit. Of course, the maximum available power from the new source 'V1new, R1new' is less than what was available from the original source 'V1old, R1old' by the amount that was dissipated in R2. From now on, V1new and R1new will be referred to as V1 and R1. The RF Source Voltage (V1) is assumed to be un modulated CW.

The transformed V1 (RMS) and R1 represent a Power Source of available power Pa = (V1^2)/(4*R1). This is the most power it can deliver to a load. It is also sometimes called the "Incident Power". For the load to absorb this power, the load itself must equal R1, and then it is called an 'Impedance Matched Load'. Changing the impedance transformation in the tuned circuit(s) changes the values of V1 and R1. This does not change the available power. That is still (V1^2)/(4*R1). As an illustration, if V1 is doubled, R1 must quadruple thus keeping the power the same.

The approach we will use in this analysis is to minimize impedance mismatch power loss between the transformed antenna resistance and the diode detector input RF resistance as well as between the detector audio output resistance and the headphones. We will show that the diode detector power Loss (DDPL), for very weak signal levels, can be minimized by using a diode with as low a Saturation Current (Is) as possible if all else is equal. In addition, the lower the ideality factor (n) of the diode, the greater will be the sensitivity to weak signals. The limitation here is that if a diode with a lower Is used, the required diode RF source and audio load resistances go up in value. That limit is reached when the diode is connected to the top (the highest impedance point) of the tuned circuit. The

the DUT in the table above. When doing a comparison of this type (comparing one element of the pair in a full headphones, to a single "standard element"), first check the volume in each of the two elements of the pair. If they are not equal, error will result. If the volumes are not too far apart, perform the measurement for each element of the pair and average the result. There is some error introduced by the procedure given above because the acoustic loading on each earphone of the pair is not the same.

Summary:
The Western Electric #509W headphones tested 6 dB less sensitive than the "standard".
The Baldwin type C headphones tested 9 dB less sensitive than the "standard".
The Brandes Superior Matched Tone headphones and the two Radio Shack Piezoelectric speakers tested 12 dB less sensitive than the "standard".
The each of two Mouser Ceramic Earpieces tested 20 dB less sensitive than the "standard".
The sound powered elements turned out to be the most sensitive and are therefore to be prefered for use when listening to weak signals, as is the case when trying for DX with a crystal radio set.

In all cases it is assumed that the source resistance driving an element is equal to the average impedance of the element over the audio frequency range of interest. This is the closest that we can get to an impedance matched condition.

Last item: Remember that headphone sensitivity can vary from unit to unit. The figures given above are not gospel for all units of a particular model. Diaphragms warp, magnets the elements is to use an audio transformer to impedance match the diode detector output resistance to the average impedance of the element. The transformer supplies supply a DC return path for the diode. A further advantage of using a transformer is that no DC voltage can get across the piezo element. Sometimes, if a strong signal is tuned in and it produces a large rectified DC voltage on the element, the element will "freeze" and its sensitivity will drop. See Article #5 for info on transformer coupling and diode DC resistance loading. (The value of the DC load resistance on the diode should equal the average value of the AC audio load impedance.)

* The comparison of the sensitivity of an element in a series connected element pair DUT with the "standard element" was made in the following manner: The full (two element) headphones DUT was connected to the J1 output of the DFLVORA. The DFLVORA was fed by a weak voice signal and the source resistance switch adjusted for the greatest volume and intelligibility. The "standard element" was then connected to the J2 output and the 3, 6, and/or 12 dB attenuators were adjusted so that the intelligibility of the voice in the "standard element" was equal to that in one element of the headphones DUT .(The other element was left dangling.) The amount of attenuation placed in the circuit feeding the standard element is a measure of the difference in sensitivity between the standard and the DUT. Since 1/2 the power going into the full headphones DUT goes into each element, one element of the DUT headphones being listened to receives 1/2 the power (3 dB less power) than that delivered to the full headphones, giving the reading for a single element a 3 dB handicap. Thus, the sensitivity of one element of the headphones DUT is 3 dB better than the sum of the readings of the attenuators. This 3 dB correction is made in the figures for

high frequency audio cutoff point may be reduced because of unavoidable winding capacitance in the audio output transformer acting against the required higher transformed headphone effective impedance.

The most important diode parameters to consider for Xtal radio operation are saturation current 'Is' and 'n'. They show up in the Shockley diode equation: Id = Is*(exp((Vd-Id*Rs)/(0.026*n)) -1), at room temperature. In crystal radio applications, the Id*Rs term may be neglected because it is usually much smaller than V. The equation then becomes:

Id=Is*(exp(Vd/(0.026*n))-1). (1)

This equation provides a good approximation of the V/I relationship for most diodes, provided the parameters Is, n, and Rs are really constant. Some diodes, especially germanium and silicon junction diodes seem to have Is and n values which increase at very high currents (higher than those usually encountered in crystal radio operation). In some of these diodes, the values of Is and n also increase at very low currents, harming weak signal reception. Is and n are usually constant in Silicon Schottky diodes, over the current range encountered in crystal radio use.

n = Ideality factor, sometimes called emission coefficient. This parameter is usually between 1.05 and 1.15 for silicon Schottky and germanium diodes commonly used in crystal radios.
Vd = Diode voltage in Volts
Id = Diode current in Amps
Is = Diode Saturation current in Amps
Rs = Diode parasitic series resistance in ohms (usually small enough to have no effect in Xtal radios)

Agilent specifies the values of Is, Rs and n for Schottky diodes in their catalog. They are listed in the table of SPICE parameters. To find some SPICE parameters for other diodes (germanium types etc.), one can use used a neat Computer Program written by Ray Waugh of Agilent. To use it one measures the diode forward voltage at five different currents (0.1 mA, 1.0 mA, 4.8 mA, 5.0 mA and 5.2 mA). Ray&39;s program runs on Mathcad 6.0 or higher. One enters the five voltages and voila, out come Is, n, and Rs. Remember this caveat: The program assumes that Is, n and Rs are constant and do not vary with diode current. If they do vary, one can change the first two currents (0.1 and 1.0 mA) to cover a smaller range, say, two-to-one, that bracket a desired diode operating current and get the Is and n values for that current. Ray told me that if anyone wants a copy of this program, it would be OK for me to supply it. A simplified method of approximating Is (n must be estimated) that does not require having Mathcad is described in article #4. A complete description of a test set-up and calculation method for determining both n and Is is shown in Article #16.

Here is what I have found experimentally through a SPICE simulation of a diode detector. If a detector diode is fed by an RF source resistance of n*0.026/Is ohms and is loaded by an audio load resistance of n*0.026/Is, then both input and output ports are matched with a return loss of better than 18 dB, assuming the signal is of weak to medium strength. This satisfies the condition of very low mismatch but only holds true for diode rectified currents of up to about 5*Is. An impedance matched diode detector insertion loss at a rectified current of 5*Is is about 3-4 dB.

Device under Test (DUT)	Sensitivity of the DUT Compared to the "Standard"	Optimum Source Resistance for the DUT	Acoustic Power Output of the DUT, compared to the "Standard" in %
Mouser #25CR035 Piezo-Electric Ceramic Earpiece. Internal capacitance=13 nF	-20 dB	12k-18k Ohms	1
Radio Shack #273-091B Piezo Speaker Element held flat against the ear. Internal Capacitance=46nF	-12	5k	6
One element of a stereo magnetic earbud that came with a small Grundig radio	-32	120 (One element)	0.06
One element of a No-name stereo magnetic earbud	-26	26 (One element)	0.25
Western Electric Model #509W Headphones (element pair connected in series)*	-6	19k (Two elements connected in series)	25
Baldwin type C Headphones with mica Diaphragms (element pair connected in series)*	-9	12k (Two elements connected in series)	13
Brandes Superior Matched Tone Headphones (element pair connected in series)*	-12	12k (Two Elements connected in Series)	6

These comparisons were made using a voice signal from a small transistor radio fed into a DFLVORA with the radio volume set to a level at which I estimated I could understand about 50% of the words. Results may differ for people who are not old and do not have poor high frequency hearing, such as myself. The differences in tone quality between the standard element and a DUT will have a different effect on intelligibility for different people.

The sensitivity values for the piezo electric elements can only be attained if the elements do not have a resistor placed across them to supply a DC path for diode detector current. The resistor adds loss (although this is a low cost approach to provide a DC path for the diode current). Also, if the resistor is made large in order to reduce its loss contribution, audio distortion will oftentimes take its place. The best way to drive

To help in understanding these charts, consider that an eight dB (6.3 times) change of power is usually perceived as a two times subjective change in loudness.

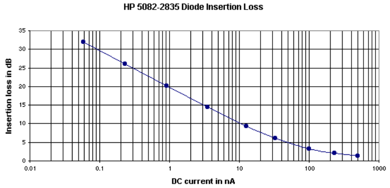
2. Comparisons

In order to compare the sensitivity of headphone elements that are used flat against the ear, as well as those that are not, (but are inserted into the ear canal (tips) or outer ear (buds)), I decided to make one of my best elements a "standard" and compare the others to it using a DFLVORA (see Article #3). That "standard" is a Western Electric Sound Powered receiver element # D173012 held flat against one ear. In the results shown below, two Mouser elements were tested and found to be of equal sensitivity. (That was after I found one Mouser to be weak until I whacked it several times.) The two Radio Shack units also tested equal. Note that the DFLVORA can be used to easily compare the power sensitivity of any two earphone elements even if they differ greatly in impedance. For more information, see the next-to-last paragraph in Section 1 of Article #2.

Table 2 - Comparison of the Sensitivity of Selected Headphones,
and Headphone Elements to a "Standard".

The input and output impedance match starts deteriorating with a DC rectified current of over about 5*Is because of the change from square law operation towards linear response at the higher input levels. At the highest RF Power input level point shown in the following graph, the rectified DC current is 500 nA and the input RF Return Loss (impedance match) is -12 dB. Diode detector power loss is 1.39 dB. At these high levels of Input Power, good matching conditions are restored if the Input Source Resistance is kept at n*0.026/Is and The Output Load Resistance is increased to 2*n*0.026/Is. If this is done input return loss goes to -26 dB and the insertion loss reduces to 0.93 dB.

Here is a graph of Diode Detector insertion power loss of an Agilent 5082-2835 or HSMS-2820 Schottky diode detector driven by a 1.182 megohm source and loaded by a 1.182 megohm load. Note that these are very high resistance values for a usual Xtal radio. The SPICE simulation was done using an Intusoft ICAP/4 simulator. Is of the diode=22 nA, n=1.03. The plot shows the insertion power loss as a function of the resultant rectified DC current.



Graph of Diode Detector Insertion Loss vs Rectified Current.

2. DISCUSSION

In general, headphones should be impedance matched by a transformer to the output resistance of the diode detector. To use a diode of such a low Is as 22 nA with, say, a Brandes Superior 12k Ohm AC impedance 2k Ohm DC resistance headphones, an impedance transformation of 1,182,000/12,000 = 98.5:1 is needed (this high a ratio is hard to get). See Article #2, "Personalized Headphone Impedance" (PHI). One should be cautious of some small (maximum dimension of less than one inch) , high transformation ratio transformers because they may have a high insertion power loss. They also may also show the effects of nonlinear inductance because the initial permeability of the core is not high enough. Their shunt inductance is usually so low at low xtal radio DX power levels, that the specified low frequency audio cut-off spec is not met. At the transformer's rated power level, the shunt inductance is generally high enough so that the low frequency cut-off spec is met. See Article #5 for info on various audio transformers.

Headphones such as the 2000 DC ohm Brandes Superior have an effective AC impedance of 12,000 ohms (PHI), but a DC resistance of 2000 ohms. If the Brandes' impedance is incorrectly considered to be 12,000 ohms at DC and audio frequencies, and is used in a 12,000 ohm circuit (without a transformer), too high a diode DC current will be drawn because the DC resistance is really 2000 ohms, not 12,000. This will load down the output RF tuned circuit thus reducing selectivity and also give increased insertion power loss. For best selectivity and minimum audio distortion at medium and

response below 500 Hz and results in a somewhat greater actual power power loss than is shown below.

4. It's a good idea to make sure that the two elements used have about the same sensitivity, otherwise the result will be the average of the good and not-so-good elements. The result will be a higher loss than if two good elements were used.

Table 1 - HPEL of some Representative Headphone elements (average of two elements).

Device under Test (DUT)	e1 in mV	e2 in mV	Optimum Source/Load Resistance for the DUT	HPEL in dB	HPEL: Acoustic Output Power vs. Available Input Power
Western Electric D173011 Sound Powered Transmitter Elements	105	12	600 Ohms	6.4	23%
Western Electric D173012 Sound Powered Receiver Elements	94	13.3	600	5.5	28
RCA/GE Sound Powered (Receiver?) Elements*	94	11.4	600	6.2	24
Brandes Superior "Matched Tone" Earphone Elements**	685	5.6	6000	18	1.6

* One of the RCA/GE units was about 6 dB less sensitive than the other. Thanks to Dieter Billinger (sky_wave_99), I knew that some RCA/GE elements having low sensitivity could be improved by sticking a small neodymium magnet to the outside of the case. It worked in this case, increasing sensitivity of the weak element by 6 dB, so it was somewhat more sensitive than the other element. BTW, a magnet could not increase the sensitivity of the other originally more sensitive element. These two units appear to be of somewhat different construction. To easily compare the power sensitivity of any two elements, even if they differ widely in impedance, see Article #3.

** Ttwo individual elements were selected for having strong magnets and their air gaps were optimized. Run-of-the-mill Brandes elements may not be as sensitive.

The I..D. of the O rings is specified as 0.987" and thickness as 0.103". 1" copper tubing is specified to have an I.D. of 1.025 and an O.D. of 1.125". The total enclosed volume is about 6 cubic cm. An alternative coupler that may give similar results is a stack of eight or nine 1/8 inch thick garden hose washers having an ID of about 5/8 inches. The ANSI 9-A coupler is (was?) the standard coupler used in Audiometry when calibrating an earphone element with a standard microphone. It is a greatly simplified version of a model of the human ear canal with an earphone cushion pressing on it.

Comments:

The DVM should preferably be an RMS responding instrument. The typical DVM responds to the full wave rectified average signal and will probably be satisfactory. Don't use a meter that responds to the peak or peak-to-peak value of the AC signal.

1. A pink-noise generator can be substituted for the white-noise generator, but it is hard to use. It has a larger low-frequency output than does the white-noise generator and therefore will show a greater fluctuation in the output as read on the DVM.

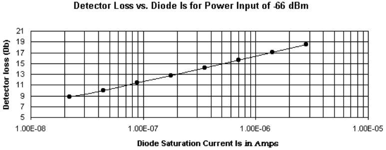
2. The noise output voltage of the white noise generator will probably have to be amplified (or increased with an audio transformer) when measuring headphone elements having a high HPEL, in order to overcome ambient noise and hum pickup.

3. The measurement method described here does not include the effect of the usual air leak between the ear pinna and the headphone element cap. This leak rolls off the low frequency

high signal levels, the DC load resistance on the diode should be the same as the AC audio load. The solution to this problem is to place in series with the headphones a parallel combination of a 10,000 ohm resistor shunted by a cap large enough to bypass the lowest audio frequency of interest. When a transformer is used; the parallel RC* (See R3 and C2 on the schematic above.) should be connected in series with the low end of the high impedance transformer primary winding. In this case the resistor should equal the transformed effective headphone impedance (PHI). Another advantage that accrues from adjusting the diode DC load to equal the AC load has to do with the way selectivity varies as a function of signal level. When the diode DC load is much smaller than the AC load (the case when using a transformer and no parallel RC), selectivity starts to reduce more and more as signal strength increases above a moderate level. The reason is that the detector rectified current increases very rapidly because of the low DC diode load resistance. A high rectified DC current always reduces the input and output resistances of a diode detector. Audio distortion may also appear. Now make the DC load higher, say equal the AC diode load impedance and have the detector impedance matched at both input and output (at low signal levels). What happens then? As the signal strength increases above a moderate level, the selectivity will change by a much smaller amount because the RF resistance of the diode detector will not drop as much as it did when the DC load resistance was small. The resistance does not drop as much because the DC rectified current is less because the DC diode load resistance has been set to a higher value than before. Impedance matched conditions also result in less power loss with consequently higher sound volume. If the headphone effective impedance over the frequency range 0.3-3.3 kHz is transformed to a value lower than the output resistance of the diode, these beneficial effects are reduced. If no transformer is used, these effects may be hard to observe because the headphone effective impedance will probably be lower than the output resistance of the diode. Also, headphones usually have a resistive impedance component about 1/6 the average value, and that goes part way towards being equal to 80% of the effective impedance.

* This may be the first time anyone has suggested placing a parallel RC in series with the diode to enable adjusting its DC load resistance equal its average AC load. Some people call it a "benny".

What is the advantage of using a diode with a low Is? We will see that if matched input and output impedance conditions are maintained, diodes with lower Is give higher crystal radio sensitivity (lower diode detector power loss) than diodes with higher Is, all else being equal. The statement above is especially important when dealing with low power signals that themselves result in high DDPL. The following graph shows the relationship between Diode Detector Power Loss at a relatively low DC Power Output Level (-66 dBm) vs. diode Is for diodes having an n of 1.03. Note that the graph data is valid only under the condition that the input and output are power matched.



Graph of Diode Detector Insertion Loss vs Saturation Current.

Figure 2 shows the circuit of a bandpass filter having -3 dB points of 0.3 and 3.36 kHz with a loss of 0.4 dB at 1.0 kHz. It's powered by a common 9 V. battery. Since the typical 1000 DC Ohms magnetic earphone element has an impedance of 6000 Ohms and the typical balanced armature sound powered magnetic element has an impedance of 600 Ohms, approximations of these two resistance values are included in the switched resistive output controlled by S1. An IC suitable for the circuit is an LF353 or an MC34002.

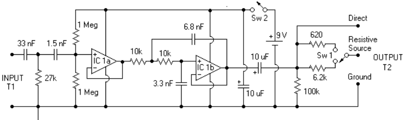


Fig. 2 0.3--3.3 kHz Bandpass Filter with Switched Output Resistances.

Schematic of Amplified Filter for use in Earphone Element Testing.

A convenient source for white noise is the headphone output of a small FM/AM transistor receiver, switched to FM and tuned to a point on the dial where it receives no signal, just noise. The noise density is rolled off at 6 dB per octave above 2.1 kHz by the de-emphasis filter in the receiver, but this should make little difference in the results. The acoustic coupler used to couple the two elements under test is based on the ANSI 9-A earphone coupler. See "Acoustic Measurements by Leo Beranek, pages 743 and 744". An approximation to the ANSI 9-A coupler can be made of a piece of 1" nominal diameter, medium weight copper tubing having a #120 O-ring glued on each end. The length of the copper tube used is 0.26 inches.

times the DC resistance of the element, assuming it is a magnetic element.

2. Couple two identical elements A and B together with an appropriate acoustic coupler and hold everything in place with several heavy rubber bands.

3. See Fig. 1. Connect a white noise generator through a 0.3-3.3 kHz bandpass filter and a source resistor Rh of value Zh to element A. Connect element B to an output load resistor Rh of value equal to Zh. The filter is necessary to limit the bandwidth of the white noise signal to the audible range of interest. If this were not done, the reading of e1 would be too high, since the noise at that point covers a wider band than that at the output.

4. We will measure the HPEL by the insertion loss method. See the section on "Maximum Available Power" in Article #0 and the Part 4 of Article #5 for information on this method. Measure the input voltage e1 at point P1 and output voltage e2 at point P2. The HPEL = 5*log (4*((e2/e1)^2)) dB. The 5 is there instead of the usual 10 because only half the measured loss can be attributed to one element, and we are actually measuring the sum of the two losses. Note: It is usually recommended that elements A and B be pressed together with a force of 1 to 2 pounds so that no air leak occurs between the elements and the coupler. Actually, if squeezing elements A and B together more tightly than the rubber bands do does not increase the value of voltage e2, the rubber bands are OK to use by themselves.

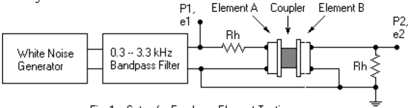


Fig. 1 -- Setup for Earphone Element Testing.

Setup for Earphone Element Testing

NOTE: There is an error in the title of the graph. It should read: Detector Loss vs. Diode Is for a DC Power Output of -66 dBm.

I used the -66 dBm signal level for the graph because it is related to the weakest voice signal I can hear with my most sensitive headphones, and still understand about 50% of the words. Here is the listening experiment that I used to determine that power level. I fed my headphones directly from a transistor radio through my FILVORA and reduced the volume until I judged I could understand about 50% of the words of a voice radio program. This enabled me to determine the average impedance of the headphones. (See article #2). I then measured the p-p audio voltage (Vpp_audio) on the headphones with an oscilloscope. Assume the AM station was running at about 100% modulation. The peak instantaneous audio voltage at the detector will be equal to Vpp_audio since the modulation is 100%. Now make the assumption that a CW carrier is driving the detector at such a level that the DC output voltage (Vdc) at the detector is equal to Vpp_audio. That DC voltage across a resistor of value equal to the detector load resistance will deliver an output power of Pdc=10*log((1000*(Vdc^2))/Rload) dBm. Since I could not get into the radio to measure the actual detector voltages and the audio load resistance, I used the p-p voltage measured across my 1200 Ohm headphones in place of Vdc to calculate the instantaneous power at the modulation peaks. Pp=10*log(1000*((Vpp_audio^2)/1200= -66 dBm. This power, Pp is that used in calculating the graph above. In my case Vpp_audio = 0.00055 Volts and effective headphone impedance = 1200 Ohms.

To calculate the actual audio power level I was using in the listening experiment, I assumed that the demodulated audio voltage was a sine wave (not a voice) with the same p-p value as the actual measured voice p-p voltage. It was then a simple matter to use the p-p voltage of the assumed audio sine wave (Vpp_audio) and the effective impedance (PHI) of the headphones to calculate the power of the audio sine wave in dBm. P=10*log ((1000*(Vpp_audio^2))/(8*PHI)) dBm. This value comes out 9 dB less than the DC power of -66 dBm. Of course there is an error here in assuming that a sine wave of a specific p-p voltage has the same RMS value as that of a broadcast voice waveform of an equal p-p value. The "Audio Cyclopedia", in an article on VU meters, states that the actual power from a voice signal is 8-10 dB less than the power from a sine wave of the same p-p voltage. I'll use 9 dB. Bottom line: The audio power from a voice voltage waveform is 18 dB less than the audio power from a sine wave voltage of p-p value equal to the p-p voltage of the voice waveform. We can now calculate that the electrical power of weakest voice audio signal I can barely understand is -66 -9 -9 = -84 dBm. This figure depends on the sensitivity of the headphones used and one's hearing acuity. I used a good sound powered headphone set in this test. My hearing acuity is pretty poor.

3. PRACTICE

Keep in mind that diodes have an unavoidable back leakage resistance. Schottky diodes generally are very good in this respect. An exception is the so-called "zero bias" detector diodes. They have very high Is values and low reverse breakdown voltages and are generally not suitable for crystal radio. Germanium and cats whisker diodes are worse than Schottkys and vary greatly. This reverse resistance increases

ARTICLE 13

How to measure the electric-to-acoustic transduction power loss of magnetic and ceramic earphone elements, with measurements of some headphone receiver elements

Quick Summary: This Article describes a device and procedure for measuring the sensitivity of earphone elements. Its purpose is to provide a quantitative method for comparing elements. Elements may be easily sorted for application to listening to weak signals, as in crystal radio sets. Actual measurements of an assortment of elements is provided.

1. Measurements

The Transduction power loss of a headphone element can be defined as the ratio of its output acoustical power to input electrical power. We will call it HPEL and express it in dB. A convenient way to measure HPEL is to use one element of a pair of identical headphone elements as a speaker and the other as a microphone, acoustically couple them together and then measure the input electrical power to the speaker element and the output electrical power from other element. Ten times the ratio of the log of the ratio of output to input power is the transduction power loss of the combination of the two elements, in dB. If the two elements are identical, the power loss of each is one-half that figure.

Here is a step by step procedure:

1. Know the average impedance, Zh of the headphone elements. If you don't know it, measure it by means of a FILVORA (See Article #2 of this series), or estimate it as 6

compared to a wavelength) is proportional to its area, among other things. If the effective location of capacitor C1 could be moved to point D, and its value kept the same, the area of the loop would be doubled to 2*A (ABDEG), thus doubling the current from its induced EMF and further increasing the directive antenna gain. Let us change the lumped representation of Co as one capacitor C1 into two capacitors, one from point B to ground and the other from point D to ground. They will each be equal to one-half of Co and both still will act as a top-tat capacitor for the vertical antenna BA, otherwise known as the lead-in. The one from D to ground will be called C2. See Fig. 4. One way to add more capacitance to ground at point D is to construct two lateral arms on the antenna. See Fig. 5. If each arm is the same length as one half the horizontal element BD (They can be of some other length.), the effective capacitance they add at point D is 2*Co/2=Cadded. The sum of the two capacitors C2 and Cadded is 1.5*Co = C4. The loop area is doubled and the current from the induced EMF is tripled (because C3 is one and one half times C1). The antenna directivity may thus be increased giving more gain in the direction in which the two effects add, and less in the opposite direction. If the directive effect can be made strong enough, a cardioid pattern should result, with a good null in the "opposite" direction. A more practical approach might be to have the two side arms droop down at an angle and be secured to ground with insulated guy wires. A single vertical wire, similar to the lead-in, might work as a substitute for the two arms if its bottom end does not get too near the ground (inverted U antenna).
Note there is no "free lunch" here. To the extent that the signal pickup is increased in the direction opposite to that to which the open end points, it is reduced in the other direction.

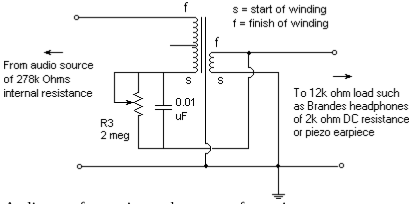
#12 Published: 07/29/00; Last revision: 11/20/05

detector loss and reduces selectivity. "n" in the diode equation is usually close to 1.05 for Schottky barrier diodes. It is about 1.15 in Germanium diodes. All diodes have a fixed parasitic series resistance Rs. It is usually low enough to be ignored in crystal radios. One problem with Schottky diodes having a low reverse breakdown voltage and low Is is that they are more vulnerable to damage from static electricity than diodes with a higher leakage resistance.

Tuned circuit loss and bandwidth considerations: A practical problem in using a diode of low Is is getting a high enough tuned circuit impedance for driving the diode. Of course, the first thing to do is to tap the diode all the way up on the output tuned circuit. An isolated tuned circuit having a typical Q of 350 at a frequency of 1.0 MHz, with a circuit capacitance of say 100 pF, and not coupled to an antenna or detector diode will have a resonant resistance of about 560k ohms. RF bandwidth will be fo/Q = 2.86 kHz. If an antenna resistance is now coupled in sufficiently to drop the resonant resistance by half to 280k, all of the available received RF power will be dissipated in the resonator, resulting in a bandwidth of 5.72 kHz (loaded Q of 175). If a diode is selected to match the now 280k ohm source resistance, it will present a 280k RF load resistance and result with a tuned circuit loaded Q of 87.5 giving an RF bandwidth of 11.4 kHz. The overall power loss caused by the tuned circuit loss is 3 dB. The diode will only receive 1/2 the maximum available-power at the antenna. The diode should have an Is of about n*0.026/278k = 100 nA (assuming a Schottky barrier diode is used). Note this: Even though the the diode is driven from a perfectly matched source (parallel connected combo of 560k tuned circuit loss and 560k antenna resistances), now the antenna does not see a matched load. It sees a parallel combo of the tuned circuit loss resistance of 560k and the 280k RF resistance of the diode.

This is a resistance of 187k ohms. This mismatch power loss, included in the 3 dB above can be partially recovered by properly and equally mismatching the antenna and the diode. If this is done by more loss-less impedance transformation (technically, with an S parameter return loss of -11.7 dB), the total tuned-circuit power loss reduces to 2.63 dB, a reduction of 0.37 dB (pretty small, but it's there). If the ratio of unloaded to loaded tuned circuit Q was less than the 4:1 ratio used here, the loss reduction would be larger.

Audio impedance transformation: One way to transform the 12k ohm effective impedance of a 2k ohm DC resistance Brandes Superior headset up to 280k ohms is to use an Antique Electronic Supply # P-T156, Stancor A53-C or similar 3:1 turns-ratio inter stage transformer. I measure an insertion power loss of only 0.5 dB with the following connection (See Articles #4 and #5 for other options.):



Audio transformer imopedance transformation

Note that the impedance transformation ratio is 16:1 thus stepping up the impedance of the 12,000 ohm headphones to 192,000 ohms not 278,000 ohms. This represents a mismatch of about 1.5:1. It will add a mismatch insertion power loss of

The first mode is that of a capacitively loaded vertical antenna. See Fig. 1. Co represents the "top hat" loading capacity. Visualize the horizontal wire BCD shifted to the left so as to be symmetrical relative to the downlead AB (with point C of ABC directly above vertical downlead BA). The total capacitance, Co, of wire BD to ground acts as the top hat capacitance for the "vertical antenna" downlead BA (the additional capacitance between the upper and lower parts of down-lead BA can be ignored because the antenna is assumed to be short compared to 1/4 wavelength). The second mode is as single turn 'virtual loop' antenna.

Let X represent a crystal radio set with antenna and ground connections, a and g. Fig. 2 shows Co as a lumped capacitor C1, connected to ground at the center of the BCD horizontal element. The rectangular loop circuit ABCFG, consisting of the four sides AB, BC, CF and FG can be looked at as a single turn loop antenna of area A, oriented to pickup signals from the B<->D direction. Note that the path for the 'displacement current' through C1 makes up one side of the loop. The current flowing from the induced EMF in the loop will combine to current from the EMF picked up in the down lead BA. These two EMFs add when a signal comes from the direction opposite to that to which the open end points and tend to cancel when coming from the opposite direction. This gives a plausible explanation of the mild directivity of the inverted L antenna. This directivity has been theoretically demonstrated by using computer simulation. For clarity, Fig. 3 shows a top view of the antenna.

The following changes to the configuration may serve to increase the directional gain of the antenna (I haven't tried them out.). The induced voltage in a loop antenna (small

ARTICLE 12

Directivity of the "Inverted" L Antenna, with Speculation as to why it Occurs and How one might enhance it

Computer modeling of the inverted L antenna shows a small directivity with the greatest signal pickup from the direction opposite that to which the open end points. I have given some thought as to why the inverted L might exhibit directional characteristics for the reception of ground (vertically polarized) waves, and present some ideas here.

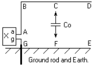


Fig. 1. Side view of the antenna-ground system, showing distributed capacitance to ground.

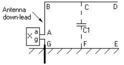


Fig. 2. Side view of the antenna-ground system, showing the antenna capacitance lumped at the center.

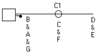


Fig. 3. Top view of the antenna-ground system.

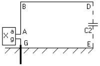


Fig. 4. Side view of the antenna-ground system, showing the effective antanna capacitance lumped at end the far end.

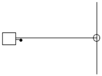


Fig. 5. top view of the antenna-ground system, showing two lateral arms.

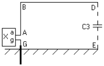


Fig. 6. Side view of the antenna-ground system, with the two lateral arms.

Figs. 1-6 show an Equivalent Circuit of the Inverted L Antenna and successive transformations thereof.

First off, understand that I am not an antenna engineer and present these speculations to suggest a way to increase the directive gain of a small (compared to a 1/4 wavelength) antenna. For simplicity of discussion, wave propagation issues will not be delved into. We will consider that the operation of an inverted L antenna results from the sum (superposition) of two modes of operation.

only 0.15 dB. If the impedance mismatch had been 2:1, the insertion power loss would have been 0.5 dB. A 4:1 mismatch gives an insertion loss of 1.9 dB.

The lead grounding the transformer lamination stack and frame is used if the transformer is mounted on an insulated material. It prevents the buildup of static charge on the frame during dry weather. Discharge of it might cause a crackling sound in the headphones or damage the diode (I got the crackling sound until I made the grounding connection).

The transformer windings start and stop leads should be connected as shown to minimize the effect of the primary to secondary winding capacitance. If the f and s connections are reversed, the capacitance between the end of the secondary and the start of the primary winding will be across the primary and reduce the high frequency cut-off point. The lower impedance (secondary) winding is usually wound on the bobbin first, then after winding on several layers of insulation film, the higher impedance (primary) is wound.

To determine how to connect the leads of the transformer, connect the primary and secondary windings as shown. (Disregarding the s and f notations). Connect an audio generator set to 1.0 kHz through a 200k ohm resistor. Load the secondary with a 12,000 Ohm resistor. Probe the input and output voltages with a scope. The output voltage should be about 0.25 of the input voltage. If the output voltage is about 0.5 that of the input, reverse the secondary leads. Repeat the test at 20,000 Hz and note the input and output voltages. Now reverse both the primary and secondary leads and repeat the 20 kHz test. The connection that gives the largest output voltage at 20 kHz is the correct one.

Note that R3 is shown above as a rheostat not a fixed resistor. The nominal setting under the low signal level conditions discussed here is about 192k Ohms. Setting it to zero has little effect on reception of these low level signals. With this design approach, when receiving high level signals, RF selectivity is not reduced as much as when the DC resistance in the diode circuit is substantially below the effective impedance of the headset. When receiving very strong signals, R3 should be set for minimum distortion.

One last comment: These design values are not critical. If impedances vary by several times from the optimum values, usually only a small sensitivity reduction will occur.

What is the effect on the volume in the headphones of a change of X dB? Many years ago I did a study which determined, in a blinded condition, that a +1.0 dB or a -1.0 dB change in sound level was barely discernible by most people. Half couldn't tell if the sound level was changed or not after being told that a change might have occurred. Another study had the listener listen to a sound. The sound was then turned off for several seconds and then on again at the same level, at a level of +3.0 dB or at a level of -3.0 dB. After the delay, only half the listeners could tell whether the level of the sound had changed or not. Incidentally, the listeners were not golden eared hi-fi listeners.

4. SUMMARY

This design approach for crystal radios provides the following benefits:

*

period is higher than when small signals are applied. Since the junction capacitance reduces when reverse bias increased, the average bias over one RF cycle is less than it is for small signals. Thus, when the signal level applied to a CSUT is increased, the frequency to which it is tuned also increases. All semiconductor diodes exhibit some of this varactor-type behaviour.

12. If the receiving antenna has a different internal resistance than the 25 Ohms used in the AMCS dummy antenna, the calculated values of S11 and VSWR and IPL will be in error. I may develop a simple way to measure the input resistance of a CSUT and will add it to this article if I do.

#11 Published: 07/21/00; Last revision: 04/10/2003

became -82.9 dBW, IPL = 4.5 dB, -3 dB RF bandwidth = 30 kHz and SF = 9.0. Tapping the antenna 2 turns from ground increased the -3 dB selectivity to 8 kHz, kept the SF at 9.0 and increased the IPL by about 4.9 dB. Note: The IPL figures use the 0.9 dB correction for the 100k resistor feeding the scope cable and also include the estimated transformer loss of 0.4 dB. A SPICE simulation of this set-up with no loss in the tank gives, for the 6-turns-from-ground tap condition, an IPL of of 6.1 + 0.4 (for the output transformer) = 6.5 dB instead of the 9.6 dB and 1.7 + 0.4 (for the transformer) = 2.1 dB instead of the 4.4 dB. This suggests a tank loss of about 2.7 dB.

10. The lower the IPL crystal radio set, the more noticeable will some of the effects noted above become.. The use of a parallel RC in the transformer primary for reducing distortion when receiving strong signals is important if the audio load resistance is higher than the output resistance of the CSUT. If the audio load resistance is lower than the output resistance of the SCUT, it becomes less important. This effect shows up in simple Xtal Sets that do not use an audio transformer. Here, the headphone impedance is usually lower than the output resistance of the Xtal Set. Also, the headphones' DC resistance, as a fraction of its AC impedance, is generally 2 or more times larger than the corresponding fraction looking into the primary of a headphone-loaded transformer. This goes part way towards equalizing the AC and DC impedance of the diode output load.

11. Here is an interesting point of information: The exact frequency to which the CSUT is tuned is a function of the input level. Reason? For small signals, the voltage across the diode is rather small, it is reverse biased for about 1/2 the RF cycle, and the average junction capacitance is close to the zero bias capacitance. When a large signal is present, the diode tends toward peak detection and is reverse biased for more than 1/2 the RF cycle. The average back voltage during this

The volume from very low strength (DX) signals is increased (less detector power loss).

*

Louder sound volume with less audio distortion when very strong signals are received.

*

Improved high signal level selectivity without changing coupling or coil taps. Less variation of selectivity with signal strength.

*

No need to tap the diode down on the output tuned circuit. Highest weak signal sensitivity is always achieved by connecting a good diode of the proper Is to the highest impedance point (assuming that the correct audio impedance transformation to the headphones is used and that the transformer has low loss).

*

Enables diodes with too high an Is to be used with strong signals without a large reduction in selectivity, by increasing R3.

Achieve the benefits by doing the following:

1.

Use a diode with an appropriate Is to impedance match the resonant resistance of the "antenna loaded RF tuned circuit" that drives the diode. See Article # 15 for new information on this.

2.

Match the audio output resistance of the diode to the effective impedance of the headphones by using a low loss audio transformer. See Article #5 for measurements on various transformers.

3.
Use a bypassed adjustable resistance in series with the cold end of the primary of audio transformer (sometimes called a "benny") to enable the diode DC load resistance to be made equal to the AC load impedance. This can be used to reduce audio distortion and improve selectivity on strong signals (compared to having R3=0) when using diodes having reasonably low excess reverse leakage (most "good" diodes). The Avago (formerly Agilent) 5082-2800 and HSMS-2800 Schottkys have high 75 volt peak inverse ratings and are not likely to overload when detecting strong signals when the "benny" is set to a high resistance. Other diodes, such as some germaniums, sometimes have enough internal leakage so that the DC load resistor (R3) can be eliminated.

#1 Published: 07/15/99; Last revision: 02/12/2006

7. Use a low noise 10X gain audio amplifier. An an improvement on this would be a single tuned band-pass amplifier tuned to 1,000 Hz. It will filter out some of the noise and hum that will probably be present.

8. Temporarily, for the tests, use an output audio transformer that transforms to a higher output resistance, along with its corresponding load resistor. Going from a 300 Ohm to a 12,000 Ohm output resistance will boost the output voltage by sqrt (12,000/3,00) = 6.3 times. I use two A.E.S. P-T157 transformers connected as shown in the first schematic in Article #5 as a variable-impedance-ratio transformer to boost the audio signal voltage. I also use it to experimentally determine if the load on the diode equals the output resistance of the diode. The switch position that gives the most output voltage is the one that provides the best match: (4, 16, 63 times ratio, or near the mean of two of the adjacent values).

9. Here are some test results with my single tuned crystal radio set that uses a 14 " square loop wound with #12 ga. solid wire for the resonator. The average parallel shunt loss resistance of the tank is 700k Ohms over the frequency range of 550-1650 kHz.. I use three Agilent 5082-2835 diodes (Is = 38 nA) in parallel for the detector and an audio transformer to convert the 700k Ohm (low signal) AC output resistance of the diode detector down to a 12,000 Ohm load resistor. I have not yet set up to measure a loop set directly, but have coupled in an external antenna connection to a tap on the tank 6 turns from ground. This, of course, loads the tank and results in a lower tank resistance than 700k Ohms. The input impedance match is very good The measured IPL at 1.0 MHz using the external antenna-ground connection is 9.65 dB at an input carrier power of -84 dBW, giving an audio output power of -102.9 dBW. The noise and hash on the scope prevented the measurement of selectivity. Measurements were then made at a carrier input power of -69.4 dBW. The output audio power

acts as sort of an automatic volume control. Unfortunately, this effect reduces the maximum volume one can get from the crystal radio set. The other is normal reverse current that increases rapidly above a certain input signal power and causes audio distortion as well as reduced volume. This effect can be observed when performing the IPL tests. For instance, in my single tuned loop set, several Agilent 5082-2835 in parallel, while very good with low signals, distort when the input carrier power at 50 % modulation gets above about -35 dBW. Several Agilent 5082-2800 or HSMS-2800 work fairly well at low signal levels but do not distort at the highest signal level I can supply. This improvement comes about because the HSMS-2800 has much less back leakage current than does the HSMS-2820 or 5082-2835 at high reverse voltages. This effect is more noticeable if the diode load resistance is above the optimum value than if below it.

3. If you use an audio transformer, don't forget to replace the R in the parallel RC with a pot and adjust it for the least audio distortion. Actually, I keep a pot in there all the time because the optimum value is usually zero for weak signals and about 1/2 the loaded RF source resistance driving the diode for strong signals.

4. For a given amount of output audio power, the output voltage is proportional to the square root of the output load resistance. This may cause a problem for those who use 300 Ohm Sound Powered Headphones (SPHP) and those who may want to make measurements at low output power levels. With the suggested starting output of 0.002 V p-p, the output power to a 1200 Ohm load (SPHP elements in series) is -94 dBW. It would be -88 dBW @ 0.002 V p-p if the SPHP elements were wired in parallel.

5. To take readings at a lower power level, there are several options to consider:

6. Use a more sensitive scope.

ARTICLE 02

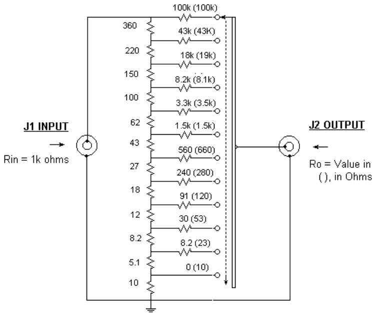
To maximize volume from headphones or a speaker, measure and use effective impedance for transformer impedance matching. No lab test equipment is needed

Quick Summary: This article describes a way to determine the effective average impedance of a pair of headphones or a speaker. This is the optimum resistance with which to drive the headphones or speaker to obtain the maximum possible volume in crystal radio set and other applications.

The magnitude of headphone or speaker impedance varies widely over the audio frequency range, being partly resistive and partly reactive. A 'Fixed Insertion Loss Variable Output Resistance Attenuator' (FILVORA) can be used to indicate the effective average value of this impedance, over that frequency range.

The first section of this Article refers to the measurement of mono headphones and individual speakers by using a FILVORA. The second section describes how to use the FILVORA to determine the effective average impedance of each element in a stereo headset. The third section describes how the FILVORA was designed.

Section 1.



Schematic of the FLVORA.

Front of the FLVORA.

demodulated output voltage Eohi_pp. Reduce the resistor to 0.7*RL and call the new output voltage Eolo_pp. Ro = 1.3*RL*((Eohi_pp - Eolo_pp)/((13/7)*Eolo_pp - Eohi_pp)). Ro varies with change of input power. At low input power levels, Ro, measured at the diode detector output (before any step-down from an audio transformer), will equal about 0.026*n/Is. At high input power levels, in the region of peak detection, Ro will approach twice the antenna-loaded RF tank resistance.

Section 6. Comments.

1. Remember that output transformer loss is included in the measurement of IPL. The usual audio transformer loss is in the range of 0.5-2 dB, but some are higher. It's a good idea to to check the loss of the one being used. A method is given in Article #5. Don't forget to reduce the calculated IPL by the 0.9 correction factor if you are using the 100k resistor in series with the scope. The MAP of the RF carrier only from the AMCS to the CSUT is: ((Ea_pp + Ea_vv)^2)/(3200) Watts.

2. It's possible for two different CSUT to have the same IPL at moderate input signal powers, but differ when receiving weak stations or strong ones. Very low input signal performance is enhanced (better DX) if the RF tank resonant resistance and transformed audio load can be made a high value. This enables the optimum diode to be one of a lower Saturation Current. The result is less IPL at low signal levels. See Article #1. Very high input signal performance is enhanced (louder maximum volume) if diode reverse leakage is kept low. This point is often overlooked. Diodes vary greatly in reverse conduction current. There are two kinds of reverse current effects: One is a gradually increasing reverse leakage current that loads the circuits more and more if the input signal increases, maybe by tuning to a stronger station. It

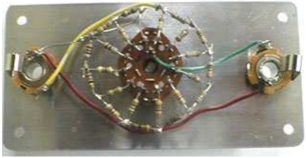
correct an audio transformer is used with a correct parallel RC in series with the cold lead of the primary of the transformer. See Article #1.

Section 4. Measurement of Input Impedance Match

Impedance Match (IM) refers to how closely the input impedance of a device equals the conjugate of the impedance of the source driving it. We will define the IM of a CSUT only at the frequency to which it is tuned. It's assumed that the input impedance is resistive at this frequency. Impedance match may be defined in terms of "Voltage Reflection Coefficient" (S11) or Voltage Standing Wave Ratio (VSWR). Either can be calculated from the voltages appearing at test points P1 and P2. Turn off the modulation of the SG. Define: RF voltage at P1=EP1_pp and voltage at P2 = EP2_pp. S11 = 20*log abs(1 - 2*(EP2_pp/EP1_pp)). VSWR = (1 + abs(1 - 2*(EP2_pp)/(EP1_pp)))/(1 - abs(1 - 2*(EP2_pp)/(EP1_pp))) These calculations define how closely the input impedance of the crystal radio set matches that of the IEEE standard dummy antenna.

Section 5. Measurement of the Output Resistance (Ro) of a Diode Detector.

The addition of a variable resistor and an ohmmeter are needed to measure the output resistance of the CSUT. Connect the SG, AMCS and scope as before. Set the fo of the SG to 1 MHz and the AM modultaion to about 50% at 1 kHz. Connect the variable resistor to the output of the CSUT and set it to the nominal audio load resistance for which the SCUT is designed. Call this value RL. Pick a moderate input power, say one that delivers an audio output power (Po) of -75 dBW to RL. An output power of -75 dBW is indicated when the 1 kHz p-p output voltage Eo_pp is: sqrt(RL*(31.6*(10^-9))). Increase the load resistor to a value 1.3*RL and call the resulting



Rear view of the FLVORA.

The circuit shown above has a fixed input resistance of 1000 ohms +/- about 5%, no matter what load is connected to the output or where the switch is set. The output resistance at any switch point is about +/- 5% of the value shown with any impedance driving the input. The insertion loss of the FILVORA is 26 dB. Standard 5% tolerance resistors are used. The use of resistors that differ by +/- 10% from the values shown should not have an appreciable impact on performance of this unit.

To use the FILVORA, connect a source of audio voice or music to the input jack J1. (I use the output jack of a transistor radio for my source.) Connect the plug of the mono headphone set or speaker to be measured to the output jack J2 of the FILVORA. Adjust the switch for the loudest volume. The correct setting indicates the effective impedance is very broad and somewhat hard to determine. Call it P2. Rotate the switch in one direction from P2 for a small reduction in volume to position P1 (generally a two positions movement), then in the other direction from P2 by two positions to P3. If the volume at P1 and P3 are the same, P2 indicates the effective impedance of the headset. If the volume at P1 and P3 is not the same, increment both the P1 and P3 settings ccw or cc by one position. When you obtain the same volume at the new P1 and P3 positions, you are done. The effective headphone impedance is the calibration indication of the switch at point P2. Sometimes equal volume settings cannot be obtained with switch settings five positions apart. If this is the case, try to get equal volume settings four positions apart. If this is done, the effective impedance is equal to the geometric mean of the settings of P1 and P3. (Take the square root of the product of the calibration readings at P1 and P3.)

The effect of source impedance on tone quality. It is interesting to note, that with magnetic elements, setting the switch to a high source resistance tends to improve the treble and reduce the bass response, compared to the response where the source matches the effective impedance of the element. Setting the switch to a low resistance does the reverse. This setting rolls off the treble, and relatively speaking, improves 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 matches the effective impedance of the element. 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 average impedance source.

Here are some practical experimental ways to vary the audio source resistance of a crystal radio set when receiving medium strength to weak 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 to one or several paralleled

these bandwidths is too small, a lower modulation frequency such as 400 Hz can be used.

The MAP of the RF carrier only from the AMCS to the CSUT is: ((Ea_pp + Ea_vv)^2)/(3200) Watts.

Section 3. Measurement of Selectivity Shape Factor:
Here is a method for measuring selectivity using the instrumentation used for measuring IPL. It is adapted from Terman's Radio Engineer's Handbook: Using a CW source, measure the frequency difference between two points that lie 3 dB down on the selectivity curve. Let us call this value S_3 kHz. Measure the frequency difference between two points that lie 20 dB down. We'll call this S_20. The input is measured at test point P1. Depending upon the input signal level chosen for this measurement, the detector may not be operated in the linear part of its operating region, but partly into its square law region. This non linearity will cause an erroneous result if the measurements are made using a constant input signal level and then measuring the output at each of the four frequencies. The correct method is to measure the input required at test point P1 to attain the specified fixed output level at each of the four frequencies. The non linearity will now be the same for all measurements and cancel out. The Shape Factor (SF), of the selectivity curve of a CSUT, at a particular RF frequency and output audio power is defined as SF = ((S_20)/(S_3)). The lower the number, the better.
Things to remember: The selectivity of a CSUT varies, depending on coupling, tap settings and frequency of measurement. It is suggested that measurements be taken at 520, 943 and 1710 kHz and any other ones where you think there might be a large variation from the average. With fixed coupling settings, the SF of a CSUT can change if the input signal power is changed. This effect can be minimized if the

modulation is present, one half of the total envelope fluctuation is caused by one side-frequency and one half by the other. Two side frequencies, each of amplitude Ea_sfpp, when added to a carrier of amplitude Ea_carpp, will cause the modulation envelope to have a maximum value of Ea_pp = Ea_carpp+2*(Ea_1sfpp). The minimum value of the envelope will be Ea_vv = Ea_carpp-(2*(Ea_1sfpp)). Define: D = (Ea_pp)-(Ea_vv) = 4*(Ea_1sfpp). Rearranging terms, we get: Ea_1sfpp = D/4. We can calculate the MAP of one side-frequency as: MAP_1sf = ((Ea_1sfpp/(2*sqrt2))^2)/(4*Ra). The first "2" changes the value of Ea_1sfpp to a peak value. The "sqrt2" changes the resultant peak value to RMS. The equation, restated, is MAP_1sf = ((Ea_1sfpp)^2) / (32* Ra). The total power in the two side frequencies is twice that in one side-frequency and is: MAP_2sf = ((Ea_ 1sfpp)^2)/(16*Ra). Now, substituting Ea_1sfpp = D/4, we get: MAP_2sf = (D^2)/(256*Ra).

The output waveform shown in Fig.3 is a sine wave Eo_pp, having a DC value of Eo_dc. The audio power it supplies to the output load Rl is: Po = ((Eo_pp/(2*sqrt2))^2)/Rl. The "2" and the "sqrt2" are needed as before to change Eo_pp from a peak to peak to an RMS value. Simplifying, Po = ((Eo_pp)^2)/8*Rl. Since IPL = 10*log (Po/MAP_2sf), and we can state the Final Result we've all been waiting for, and it is:

INSERTION POWER LOSS = IPL = 10*log {32*Ra*[(Eo_pp/D)^2]/Rl}.

There is one caveat to using this method: It is assumed that the audio bandwidth through the audio transformer, as well as one half the -3 dB bandwidth of the RF tank is 3 or more times as large as the recommended 1000 Hz modulation frequency. If both are 3000 Hz, the error will be about 0.6 dB. I either of

Schottky diodes such as the Agilent 5082-2835. Schottky diodes described as "zero bias detectors' have a high saturation current and are not suitable. 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 capacitance will also reduce the treble response. A side benefit from this change, on some crystal radio sets is an increase of selectivity. This is because the RF load resistance presented to the tank is raised when the diode saturation current value is reduced.

*

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. If they are in parallel, series connecting them has the effect of decreasing the effective source resistance.

*

Refer to Articles #0, #3,#5 and #14 for more info. Consider the 'Ulti-Match' by Steve Bringhurst at: http://www.crystalradio.net/ Check out Index>Sound powered>Impedance matching>Impdance matching...

If you are interested in DX reception with headphones and do not have normal hearing, you might want to customize the source resistance driving the headphones. This enables using the 'change in headphone frequency response as a function of headphone driving resistance' to partially compensate for high frequency hearing loss. Input a voice signal and reduce its volume to a sufficiently low level such that you judge you understand about 50% of the words. Readjust the switch to see if you can obtain greater intelligibility at another setting. If you can, this new switch setting indicates the source resistance with which to drive the particular headphones being used to deliver maximum voice intelligibility for your ears. I call this resistance: Personalized Headphone Impedance (PHI). For magnetic headphones, this resistance is higher than the average impedance of the earphones, for piezo-electric ceramic earpieces, the resistance is lower.

Two FILVORA units enable one to compare the actual power sensitivity of two headphones, even if the effective impedance the two headphones are very different. A dual unit to do this (DFILVORA) is described in Article #3.

Section 2.

The effective impedance of hi-fi stereo headphones may be checked with the FILVORA. The effective impedance of the two earpiece elements can be checked by determining the switch position for maximum volume with one of these connections: (1) The sleeve, to the ring and tip in parallel or (2) the ring to the tip. Measurement (1) will show one half the effective impedance of one earpiece and measurement (2) will give a reading of two times the effective impedance of one element.

Section 3.

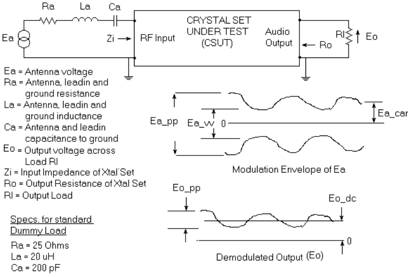


Fig.3 Equivalent Circuit of Antenna, Crystal Set and Output Load.

Equivelent Circuit of Antenna, Crystal Set and Output Load

The IPL of any crystal radio set depends upon the output power level at which it is operating. At very low output levels (signal barely readable with sensitive headphones), the IPL increases about 6 dB for every 6 dB reduction in input power. This results in a 12 dB reduction in output power. When this happens, the diode detector is said to be operating in its "square law region". Because of this effect, it is suggested that the IPL be measured at several audio output power levels when characterizing a crystal radio set, maybe -80 and -110 dBw.

Section 2. Derivation of IPL.

Figure 3 shows of the envelope of an AM carrier of frequency fo, modulated at 50% by a sine wave of frequency fmod. This modulation produces two side frequencies separated from the carrier by fmod. One is above fo and one is below it. If no side frequencies were present, Ea_pp would equal Ea_vv and the modulation envelope would be straight lines. With some

Keep the leads short to minimize 60 Hz hum pick-up. Only use the 100k resistor when measuring the output at 1000 Hz. Don't use it when measuring RF at the input. When calculating IPL, correct your results for the 0.87 dB loss (Use 0.9 dB).

* The output sine wave may look distorted. This can come from modulation distortion in the signal generator or distortion generated in the CSUT. Generator distortion is not very important here. Distortion generated in the CSUT can be caused by an incorrect resistance in the parallel RC used in series with the audio transformer primary (if one is used). To check, replace the resistor with a pot and adjust it for minimum distortion. BTW, this is the best way to find the correct value for the resistor. See Article #1 of this series.

Here are the labeling conventions that will be used. Voltages on the input side of the CSUT always start with Ea. Voltages on the output side start with Eo. The underscore is a separator from the description suffix that follows. fo = carrier frequency, fmod = modulation frequency, pp = peak-to-peak, vv = valley-to-valley, car = carrier, dc = direct current, sf = side-frequency, carpp = carrier peak-to-peak, 1sfpp = one side-frequency peak-to-peak, 1sf = one side-frequency, 2sf = two side-frequencies.

To help define the equations used to calculate the the resistor values for the asymmetrical attenuator 'FILVORA' the following requirements were set up:

1. Output resistance range: 10 to 100k ohms, with 12 switch positions. This range covers the span of impedances found in headphones used in transistor radios, up to that found in piezo earpieces. A 12 position rotary switch was used because it is readily available.

2. The FILVORA must exhibit the minimum possible constant insertion loss and input resistance, independent of switch position or load impedance selected. The maximum and minimum output resistances are 100,000 and 10 ohms. For a constant, and minimum insertion loss at all switch positions, this requires the input resistance be: sqrt(100,000*10)=1,000 ohms at each switch position. The requirement for a constant input resistance is closely met by an equation equating the sum of the 12 resistors in the vertical string equal to 1000 ohms.

3. The ratio of the output resistance from one switch point to the next shall be constant. The output resistance ratio between adjacent switch positions is (100,000/10)^(1/11)=2.3101. 12 simultaneous equations are necessary to meet this requirement on each switch position.

4. The same insertion loss shall exist on each switch position. The voltage ratio (loaded output to input) required at each switch point, for constant insertion loss (26 dB), requires another 12 simultaneous equations.

A system of 25 simultaneous was written and solved in MathCad for the values of the 24 resistors. Those are the values (5% resistor series) shown in the schematic. To minimize power loss, the attenuator becomes an inverted L minimum-loss pad at the two extreme switch positions. It is a non-minimum loss T pad at the intermediate positions. The output resistance range of the FILVORA is 10,000 to 1. This establishes the minimum loss. If the output resistance range were 100,000 to 1, the insertion loss would have to be 31 dB. Insertion power loss = 5*log(resistance ratio)+6dB.

#2 Published: 07/15/99; Last revision: 06/21/2003

value between 300 and 3,300 Hz. Also, transformer loss and distortion is usually greater at 400 Hz than at 1000 Hz.

* The usually specified modulation percentage is 30 %. I suggest using 50 %. This gives a greater output voltage and makes low signal level measurements easier.

* This test procedure uses one scope at several input attenuator settings as well as at 1000 Hz and 1.0 kHz. It depends upon calibration accuracy from one switch position to another as well as from 1.0 kHz to 1.0 Mhz. I got caught on this. My scope is 21 yeas old and the frequency response flattening trimmers in the vertical attenuator had drifted. This didn't affect the accuracy at low frequencies, but produced error at 1.0 Mhz. The best way to check for this problem is to use a quality, fast rise-time Square Wave Generator and check for a good clean corner at the leading edge of a 100 kHz square wave. Another option is to use a sine wave Function Generator, the output of which is known to be flat vs. frequency. If it has an output up to over 10 Mhz, the output is probably flat from 1.0 kHz through 1.0 Mhz.

* One probably will find an undue amount of noise, hash and carrier RF on the scope screen when measuring the output waveform. This can be caused by capacity coupling in the transformer between the hot end of the primary winding and the hot end of the secondary. I eliminate this hash by using a very simple low-pass filter. To do this, connect a 100k Ohm resistor in series with the scope input cable, very close to where it connects to the transformer output terminal. Assume that the scope has a one Megohm (check it!) input resistance, in parallel with a 175 pF input capacity when using the probe at the X1 setting. (These are the values for my Tektronix model T922 scope.) At 1000 Hz the voltage divider from the series 100k Ohm resistor and the input impedance of the probe causes the scope to read 0.87 dB less than the actual output of the CSUT. At 1.0 MHz the attenuation will be will be 41 dB.

attenuators, to hold a +/- 0.33 dB accuracy are: 3 dB-10%, 10 dB-4% and 20 dB- 2.5%.

The load on the CSUT must be a resistor (R1) of value equal to the effective impedance of the headphones used with the crystal radio set. One can determine the impedance of the headphones by building and using the FLVORA described in Article # 2. Alternatively, it may be estimated as 5 or 6 times the DC resistance of the phones.

To measure the IPL of a crystal radio set, connect the SG**** to the AMCS and set it to a test frequency of, say, 1.0 MHz. Turn on the sine wave modulation function and adjust the frequency to 1000 Hz**** and the modulation percentage to 50%****. (50% modulation exists when Ea_pp is three times Ea_vv.) Connect the AMCS to the antenna and ground terminals of the CSUT. Connect the scope to Rl and set it to a sensitivity of 2 mV/div. Set the SG to a high RF output and tune the CSUT to maximize the 1000 Hz trace on the scope****. Reduce the SG output as necessary to keep the scope trace on scale****. Reset the SG to deliver a 4 mV p-p trace on the scope. Connect the scope to point P1 and measure and record Ea_pp and Ea_vv at Point P1.
****

* Some RF signal generators have too much harmonic waveform distortion in their output to give accurate results with this procedure and will need a simple harmonic filter to clean up the output. If the RF waveform looks like a fairly good sine wave it's OK.

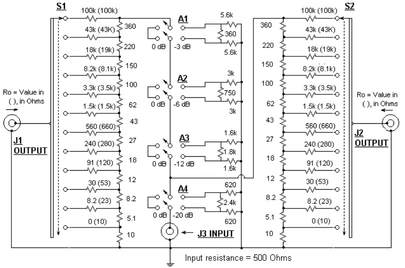
* 1000 Hz is chosen instead of the usually specified 400 Hz because most high performance crystal radio sets use an audio transformer to drive the headphones. At 400 Hz, the impedance of most transformers is well below the average

ARTICLE 03

Compare impedance and sensitivity of headphones, and/or speakers even if they differ greatly in impedance

The purpose of this article is to show how to compare the sensitivity of two pair of speakers or mono headphones even if they differ greatly in effective impedance. See Article #2 on how to measure impedance. The Dual Fixed Insertion Loss Variable Output Resistance Attenuator (DFILVORA) will be described and directions for its use will be given in Section 1. It is essentially a combination of two FILVORA units along with some extra attenuators. Section 2 will describe how to modify the DFILVORA for use with Hi-Fi stereo headphones. Note: The use of resistors that difer by +/- 10% from the values shown in the schematic should not have an appreciable impact on performance of this unit.

Section 1.



Schematic of the DFLVORA.

Connect an audio source to jack J3. (I use the output jack of a transistor radio for my source.) Connect the plug of one of the two mono headphone sets or speakers to jack J1 and the other to jack J2. Set attenuators A1, A2, A3, and A4 to 0 dB. Switch S1 should be set to the position providing the loudest sound in the headphone set or speaker connected to J1. Switch S2 should be set to the position providing the loudest sound in the unit connected to J2. (Read Article #2 to see the recommended procedure for doing this.) If the unit connected to J2 is louder than the unit connected to J1, reverse the units. Now add attenuation in the path to the unit connected to J1 in 3 dB steps by using A1, A2 and A3 in the proper combinations (the dB's add), until the volume from the unit connected to J1 equals the volume in the unit connected to J2 as closely as possible. If the sound cannot be reduced to a low enough level because of volume control limitations in the INPUT source, use A4 to reduce the volume by 20 dB.

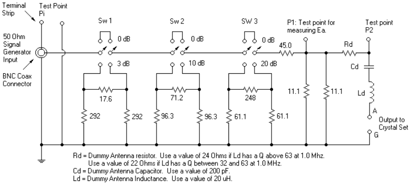


Fig. 2–Apparatus for use when Measuring Crystal Set Insertion Power Loss, Selectivity and Input Impedance. (AMCS)

Fig. 2-Schematic of an Apparatus for use when Measuring Crystal Set Insertion Power Loss, Selectivity and Input Impedance.

We will use use a special attenuator box between the SG and the CSUT and call it the AMCS. Refer to the schematic in Fig. 2. The AMCS has one 3 dB and one 20 dB attenuator that are used in measuring selectivity. It has an additional 10 dB attenuator in the event some extra attenuation is needed. The 20 dB attenuator can also be used to determine the voltage Ea at test point P1 when it is so low that it is hard to read. The series 45.0 and two parallel 11.1 Ohm resistors form a "minimum-loss impedance transforming attenuator". Its input design resistance is 50 Ohms. Its attenuation is set so that the ratio of the voltage at test point Pi to that at P1 is 10:1 when Sw1, Sw2 and Sw3 are set to zero dB. The source resistance feeding the Dummy Antenna and crystal radio set series combination is 5.25 Ohms. Two 11.1 Ohm resistors are used in place of one of 5.55 Ohm resistor because resistors under 10 Ohms may be hard to find. This also minimizes lead inductance. If the 45.0 and 11.1 Ohm resistors are held to within +/- 4%, the attenuation accuracy will be within +/- 0.33 dB. of nominal. Resistor accuracy tolerances for the other components Ra, La and Ca are intended to have the same impedance as the average antenna that used to be used for AM reception in the USA. These components are termed a "dummy antenna" and are specified for use in standardized testing of AM receivers. The standard is described in "Standards on Radio Receivers", Institute of Radio Engineers (predecessor of the IEEE), New York, 1938.

It is assumed that by tuning the CSUT for maximum output volume, that the best conjugate impedance match possible is presented to the antenna. In simpler terms, tuning for maximum volume adjusts the resistance component of Zi to as close to 25 Ohms as possible and the reactive component of Zi to as close a value as possible to the negative of the reactance of La and Ca in series. This set of circumstances transfers the most signal power possible from the antenna to the CSUT.

The test procedure that follows involves applying a modulated RF Voltage (Ea) through a dummy antenna to the input of the CSUT and then measuring the Audio Output Power (Po) delivered to the output load. The IPL of the CSUT is calculated as: IPL = 10*log (Po/(MAP in the sidebands of Ea)).

Measurement of IPL

The sum of the dB settings of A1, A2 and A3 equals the difference in power sensitivity between the two headphones or speakers, independent of the effective impedance of the units. The settings of S1 and S2 indicate the average impedance of the two units. See Article #2 for more details on this subject. The power insertion loss from jack J3 to either jack J1 or jack J2, with the attenuators set to zero, is 29dB.

Section 2.

To compare the power sensitivity of two stereo headphones, I recommend that several modifications be made to the DIFLVORA. Jack J1 should be changed to a stereo jack and the ring and tip connections tied together. The same change should be made to J2. this will cause each stereo headphone to be tested in mono mode with its elements connected in parallel. The value of all resistors should be halved. The setting of switches S1 and S2 indicate the average impedance over the audio frequency range of two earphone elements in parallel, and that figure will be one-half the value of each element by itself. Also, the measurable range of individual element impedances will be changed from 10 to 100k Ohms to 20 to 200k. The range of impedance measurable for the parallel combo of two elements is still 10 to 100k Ohms. To correct this condition and have the DFILVORA switches S1 and S2 indicate the value of the impedance of one element (and not two in parallel), halve the value of all resistors in the schematic. If this modified DFILVORA is now used to measure a mono headset, the resistance readings of switches S1 and S2 will be twice the actual value. These modifications change the input resistance of the DLVORA to 250 Ohms.

#3 Published: 07/15/99; Last revision: 11/04/00

ARTICLE 04

The best diode and audio transformer for a crystal set, and a way to measure diode saturation current

Here is a practical way to determine the diode and audio output transformer impedance matching characteristics needed to maximize sensitivity and selectivity for weak signals and to reduce strong-signal audio distortion in a Crystal Radio Set. Unfortunately, this may be an iterative process.

1. Determine the RF output resistance at resonance of the tuned circuit driving the diode while the Crystal Radio Set is connected to its antenna.
2. Calculate the Saturation current (Is) that the diode should have. The ideality factor of the diode should be as low as possible. Get an appropriate diode.
3. Know the effective impedance of the headphones to be used.
4. Calculate the impedance transformation ratio needed to transform the diode audio output impedance to that of the headphones.
5. Connect it all up.

1. Connect the Crystal Radio Set as presently configured to antenna, ground and headphones. Select a frequency for optimization. About 1 MHz is suggested. Tune the Crystal Radio Set and adjust the antenna coupling and diode tap (if there is one) for the desired compromise between sensitivity and selectivity on a signal near 1 MHz. Replace the headphones with a 10 Meg resistor load bypassed with about 0.002 uF capacitor (no transformer yet). We will now use the diode as a voltage detector. Measure the detected DC voltage with a high impedance (10 Megohm) DVM. If the diode can

of Za Ohms and the load has an impedance of Rl. The SCUT is tuned and adjusted to deliver maximum audio power to the load, with the desired selectivity. Define the output power as Po. Now imagine the replacement of the CSUT with an ILCS. It provides a perfect impedance match to the source and perfectly matches the load. Its output power will equal Pa because there are no losses. This ideal crystal radio set will function as a device to convert all of the MASP into audio power. The ratio of the output power of the CSUT to that of the ICS set is Po/Pa. This ratio, expressed in dB is the IPL of the CSUT. IPL = 10*log (Po/Pa) dB. The load resistor should have a value equal to the average impedance of the headphones to be later used with the CSUT. (See Article #2 on how to measure headphone impedance.)

Section 1. IPL Measurement Method.
The test equipment required is:

1. An RF signal generator (SG) covering 530 -1700 kHz and capable of linear amplitude modulation up to 50%. The generator can be a modern function generator or a conventional RF signal generator, provided that the RF waveform has a reasonably low harmonic content. It should have a 50 Ohm output resistance.
2. A scope with a flat response to at least 1.7 MHz and an accurate calibrated vertical sensitivity of 0.002 V per division or better. Input resistance is assumed to be 1 Megohm. Input capacitance (including that of the connection cable) is assumed to be about 175 pF.
3. A special attenuator set up and impedance adjuster unit called AMCS.

The signal source is modeled as a voltage source Ea with series internal impedance elements of Ra, La and Ca. See Fig. 3 The

Here is how to calculate MAP from the Vs and Rs combination. As mentioned before, the maximum power output occurs if Rl = Rs. This means that the total load on Vs is the series combination Rs + Rl = 2*Rs. Since power in a resistor can be calculated as (V^2)/R, the total power dissipated in the two resistors is (Vs^2)/(2*Rs). Since one half of the power is dissipated in Rs (and lost) and one half in Rl, the maximum power deliverable to Rl is: (Vs^2)/(2*Rs)/2 = (Vs^2)/(4*Rs). We will use this relationship later on. Note that Vs^2 means Vs squared and 4*Rs means 4 times Rs. Vs is in RMS volts. If Vs is given in peak or peak-to-peak units, a correction factor must be applied.

Definition of IPL when the input signal is an RF carrier, modulated by a sine wave.

Input Power: Audio information that is amplitude modulated on an RF carrier is contained solely in what are called sidebands. Sidebands are better called side frequencies if the audio modulation waveform is a single sine wave, as will be the case here. In sine wave AM, two side frequencies are generated in the modulator. One is at a frequency above the carrier and one is below it. They are each spaced away from the carrier by an amount equal to the modulation frequency. These two side frequencies carry all the information that is in the signal. The RF carrier carries none. When we receive a signal on our crystal radio set it is this sideband power that we want of capture and convert to audio power in our headphones. The carrier acts only as a "carrier" for the sidebands and generates the DC diode current and DC voltage across the DC resistance component of the load.

Output power and IPL: Assume that an RF source with a MASP of Pa Watts is connected to a CSUT and that the CSUT feeds a load resistor. The source has an internal RF impedance

be tapped down lower on the RF tuned circuit, do so until the detected voltage is as low as can easily be read. Trim the tuned circuit tuning if necessary. Find the value of a 0.125 or 0.25 watt carbon or metal film resistor which when connected across the RF tuned circuit reduces the detected voltage to about 0.35 of its previous value (retune as needed). Use short leads on the resistor. The value of the resistor (lets call it Rr) approximates the resonant resistance of the tuned circuit with antenna connected. See Part 11 of Article #0 for more info on resistor types.

What we have done here is to minimize loading on the tuned circuit from the diode detector. If this diode loading is made negligible, using a resistor of value equal to that of the resonant resistance of the tuned circuit will reduce the RF voltage to 0.5 of what it was before the resistor was placed. Here, the diode has been given a high resistance DC load to further reduce its loading effect on the tuned circuit (the 10 Meg resistor connected in place of the headset). The detector is used as an indicator of the RF voltage across the tank circuit. The diode will be operating somewhere between linear and square law. That is where the 0.35 comes from (geometric mean of 0.5 and 0.25). A Better approach, if one has a high sensitivity scope good to above 1.0 MHz, is to disconnect the diode from the tuned circuit. Then very lightly capacitively couple the scope to the tuned circuit and use it as a measuring tool when placing the resistor across the tuned circuit. Then of course, one would use the 0.5 figure for voltage reduction since the measurement is linear. Bear with the problem of the measured voltages jiggling up and down due to modulation. Just estimate an average. (See Article #0 for information on diode Saturation Current and Ideality Factor.)

2. A good diode to use in the crystal radio set above, for weak signal reception, is one with an axis-crossing resistance equal to Rr. A diode that has an axis-crossing resistance of Rr is one having a Saturation Current of Is = (25,700,000*n)/Rr nanoAmps. The ideality factor of the diode (n) is an important parameter in determining very weak signal sensitivity. If all other diode parameters are kept the same, the weak signal input and output resistances of a diode detector are directly proportional to the value of n. Assume a diode with a value of n equal to oldn is replaced with an identical diode, except that it has an n of newn; and the input and output impedances are re-matched. The result will be a detector insertion loss change of: 10*log(oldn/newn) dB. That is, a doubling of n will result in a 3 dB drop in power output, assuming the input power is kept the same and impedances are re-matched. This illustration shows the importance of a low value for n. Back leakage resistance should be low and the diode series resistance (Rs) should also be fairly low. Diode barrier capacitance should be fairly low (6 pF or less) . Schottky barrier diodes usually have low series resistance, barrier capacitance, Ideality Factor and very low back leakage. The challenge is to get a diode reasonably close to the correct Is. (If it's within 0.3 and 3 times the calculated value, you won't notice much difference.) A simple way to check for back-leakage is to measure the back resistance of the diode with a non-electronic VOM such as the Triplett 630 or Weston 980. Use the 1000X resistance switch position. If no deflection of the meter can be seen, the diode back leakage is probably OK. Another way is to place a DC blocking capacitor in series with the diode. If the audio becomes very distorted, the diode leakage is low (this is the desired result). A value of 1000 pF or so is OK for this test.

perfect impedance match to the signal source nor perfectly match the output audio load, thus incurring mismatch losses. It will have some internal power losses. Its output audio power will be less than that of the ILCS. The IPL of the CSUT is: IPL = 10*log ((Output power of CSUT)/(Output power of ILCS)) dB.

Now, a brief detour to explain the concept of MAP (Maximum Available Power) and a more detailed look at Insertion Power Loss (IPL) as used in this Article.

Maximum Available Power (MAP)

Assume that any electrical source of power can be represented as a voltage source (Vs) that has an inaccessible internal impedance Zs = Rs + jXs. See Fig. 1. Assume that the reactance component (Xs) of this impedance is tuned out. The crystal radio set tuner should do this by generating a series reactance whose value is the negative of Xs. There is a maximum amount of power that Vs, with its internal series resistance Rs, can deliver to any load. The value of the load (Rl) for maximum power transfer is Rs itself. This is called an impedance matched condition. Any other value for Rl will absorb less power from the source than a value of Rs.

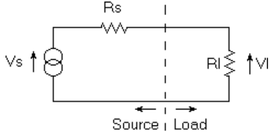


Fig.1 Power Source and Load.

Fig. 1-Schematic of a Power Source and a Load

comments and suggestions on how to improve crystal radio set performance.

A quick definition: The IPL of a crystal radio set may be loosely defined as 10 times the log of the ratio of the audio power delivered to the output load divided by the maximum RF sideband power available from the antenna.

Here comes a more rigorous definition of IPL: The function of a crystal radio set is to convert (demodulate) the modulated RF signal sideband power received by the antenna-ground system and deliver as much of that power as possible to the output load as audio output power. Understand that all of the signal information modulated on a carrier and picked up by an antenna-ground system resides in the power carried in the sidebands of that signal. No signal information is contained in the RF carrier. The Insertion Loss Method assumes that a voltage source with a specific internal impedance is connected through a "device under test" (DUT) to a load resistor. We can say that the DUT is "inserted" between the source and the load.

* First, consider what happens when an Ideal Lossless crystal radio set (ILCS) is inserted as the DUT, tuned to the source signal and adjusted for maximum output. It is connected between the signal source and a Load Resistor (Rl) representing the average impedance of the headphones to be used later. This ILCS presents a perfect impedance match to the signal source and also to the output load. It has no internal power losses. The ILCS will convert all of the Maximum Available Sideband Power (MASP) in the modulated signal source into useful audio power in the output load. Power loss is zero when the ILCS is inserted as the DUT.

* Second, consider what happens when a real world crystal radio set is inserted as the DUT. It probably will not present a

Here is an easy way to determine the approximate Is of a diode. Forward bias the diode at about 1.0 uA. A series combination of a 1.5 volt battery and a 1.5 Meg resistor, connected across the diode will do this. Measure the voltage developed across the diode with a DVM having a 10 Meg input resistance. Calculate Is=667*(Vb-Vd)/(e^(Vd/(0.0257*n))-1) nA. e = base of the natural logarithms = approx. 2.718, ^ = "raise the preceding number to the power of the following number", Vb = voltage of the battery, Vd = voltage across diode and n = diode Ideality Factor (Emission Coefficient). I suggest using an estimate of 1.12 for n. Most good detector diodes seem to have a n between 1.05 and 1.2 A method for measuring both n and Is is shown in Article #16. Measurements on 1N34A germanium diodes at various currents show that the values for Is and n are not really constant, but vary as a function of diode current. Is can increase up to five times its value at low currents when currents as high as 400 times Is are applied. However, germanium diodes I have tested exhibit a fairly constant n and Is when measured at currents below about six times their Is. A rectified current of about 6 times Is corresponds to a fairly weak signal. The following chart shows some results from measuring several diodes at a current of 1.0 uA. The calculated low-signal-level value of the diode junction resistance Rj= 0.0257*n/Is is is also given. Note the wide variation among the various diodes sold as 1N34A. Schottky diodes, as a rule are fairly consistent from unit-to-unit. The Agilent '2835 measured 11 nA, and many others test close to this value. I think that many years ago early production '2835 diodes probably matched the Spec. sheet value of 22 nA for Is. Over the years, I would guess that the average value was allowed to drift in order to optimize other more important parameters (for most applications) such as reverse breakdown voltage. BTW, Is is not a guaranteed 100% tested production spec.

Caution: If one uses a DVM to measure the forward voltage of a diode operating at a low 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, leaving only the diode in parallel with the internal resistance of the source connected to the terminals of the DVM. If the DVM reads more than a tenth or so of a millivolt, 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. Connect the capacitor across the diode with very short leads, or this fix may not work.

If one wishes to screen a group of diodes to find one having a specific Is, use the setup described above. Substitute the desired value of Is into the following equation: Vd=0.0282*ln(667*(Vb-Vd)/Is+1) volts. 'ln' means natural base logarithm and Is is in nA. A diode having a Vd equal to the calculated value will have approximately the desired Is.

Here are some tips to consider when measuring diodes: Keep all leads short and away from 60 Hz power wires to minimize AC and electrostatic DC pickup. Place a grounded aluminum sheet on the workbench, and under the DVM and other components to further reduce spurious pickup by the wiring. A piece of grounded kitchen aluminum foil will do nicely for the aluminum sheet. You may find that the reading of Vd slowly drifts upwards. Wait it out. What you are observing is the temperature sensitivity of Vd to heat picked up from

Is	Saturation Current of a diode. See Article #1 for an explanation of this term.
MAP	Maximum Available Power, in Watts.
MASP	Maximum Available Sideband Power, in Watts.
n	Ideality factor of a diode. See Article #1 for an explanation of this term.
p-p	Peak to peak.
Po	Detector Output Power, in Watts.
sqrt	Square root of the expression that follows.
RL	Detector load resistor.
Ro	Detector Internal Output Resistance.
S-3	Frequency difference between two points 3 dB down on the selectivity curve.
S_20	Frequency difference between two points 20 dB down on the selectivity curve.
S11	Voltage Reflection Coefficient.
Suffix	See the paragraph above Fig. 3 for the suffix labeling conventions used when measuring IPL.
SF	Shape factor, the ratio of the 20 dB down bandwidth to the 3 dB down bandwidth.
SG	Signal Generator.
SPHP	Sound Powered Headphones.
SPICE	A computer program used to simulate the physical operation of circuits.
Vs	RF Voltage source
VSWR	Voltage Standing Wave Ratio.

This article is divided into six sections. The first describes the IPL (Insertion Power Loss) measurement method. The second gives a theoretical derivation. The third shows a method for the measurement of selectivity. The fourth shows how to measure the input impedance match of a CSUT (crystal radio set Under Test). The fifth shows a method for measuring the output resistance of a crystal radio set. The sixth gives some

ARTICLE 11

A Procedure for Measuring the Sensitivity (Insertion Power Loss), Selectivity and Input/Output Impedance of a Crystal Radio

Quick Summary: This Article describes a device and procedure for quantifying several characteristics of crystal radio sets. They are: (1) Insertion power loss, (2) Selectivity, (3) RF input impedance match and (4) Audio output resistance.

First, an acknowledgement: This article was inspired by a paper written on 9/15/99 by Charlie Lauter at: Lautron@aol.com . It can be accessed at: http://home.t-online.de/home/gollum/testing.htm . He led the way with a good procedure for sensitivity and selectivity measurement, but I wanted a more general approach. Here is mine:

Definitions and Acronyms used in this Article

abs	Absolute value (sets the next expression to a positive value).
AMCS	Apparatus for use when Measuring crystal radio set Insertion Power Loss and Selectivity.
CSUT	crystal radio set under Test.
D	Difference between RF envelope peak-to-peak and valley-to-valley voltage.
DUT	Device Under Test.
Eo_pp	Peak to peak demodulated output voltage
FLVORA	Fixed Loss, Variable Output Resistance, Attenuator.
ILCS	Ideal Lossless crystal radio set.
IM	Impedance Match.
IPL	Insertion Power Loss in dB.

handling the diode with your fingers. Let the diode return to room temperature before taking data.

Many glass diodes exhibit a photoelectric effect that can cause measurement error. Guard against it by checking to see if a diode current reading changes when the light falling on the diode is changed.

Saturation Current (Is) and the related Junction Resistance (Axis Crossing Resistance), Rj, of some Diodes, Measured at 1.0 uA. (*=Mfg's data)

Type of Diode	Is in nA	Junction Resistance in ohms (Axis-crossing resistance)
Agilent 5082-2835	11	2.5 Meg
Agilent HBAT-5400	100*	282k*
Agilent HSMS-2870	140*	191k*
Radio Shack 1N34A (marked 12101)	160	170k
Radio Shack 1N34A (blue body marked BKC)	180	150k
Radio Shack 1N34A (brown, orange and white bands)	200	130k
Radio Shack 1N34A (labeled BKC 2000)	400	65k
Radio Shack 1N34A (clear glass)	600	45k

2N404A connected as a diode (collector and base tied together)	1700	16k

Published SPICE Parameters for some Agilent (formerly Hewlett-Packard) Schottky Barrier diodes:			
HSMS-2800 This is a SMD (Surface Mount Diode)	n=1.08	Is=30 nA	Rs=30 Ohms
HSMS-2810 This is an SMD type	n=1.08	Is=4.8	Rs=10
HSMS-2820 This is an SMD type	n=1.08	Is=22	Rs=6
HSMS-2860 This is an SMD type	n=1.10	Is=38	Rs=5.5
HBAT-5400 This is an SMD type	n=1.0	Is=100	Rs=2.4
HSMS-2870 This is an SMD type	n=1.04	Is=140	Rs=0.65
5082-2835 This is a glass type, but expensive now	n=1.08	Is=22	Rs=5

Note that these values for Is and n are not cast in stone. Is can easily vary by 2:1 or more from diode to diode of the same type.

Multiple similar diodes may be paralleled to increase Is. Is is increased proportionally to the number of diodes in parallel.

DDIPL (Lower values for attenuators A1 and A2) and greater interaction. If DDIPL approaches zero, the output resistance will approach two times the source resistance R1. Similarly, the input resistance will approach 1/2 the load resistance R2. If the input signal power is reduced, detector input and output resistance values become decoupled from each other and both approach Ro. See the paragraph below Fig. 1.

Overview: One can think of a diode detector circuit as a device to change input RF power to an almost equal amount of DC output power, provided the input power level itself is high enough. In this instance the attenuators A1 and A2 in Fig. 2 have very low values. If the input power is reduced, A1 and A2 increase in loss, thus reducing the output power. At low input power levels, square law operation occurs. In this region, if the input power is reduced by, say, 1 dB, the loss in attenuators A1 and A2 are each increased resulting in an output reduced by 0.5 dB. Voila, square law operation! There is an extra loss besides that of A1 and A2. It is the interface mismatch loss between each attenuator and the diode Di as well as input and output mismatch losses. This interface loss varies as a function of input power. It is about zero when the values of A1 and A2 are very low (large signal power condition) and approaches 3 dB for each attenuator at low signal power levels (total of about 6 dB). See the Table 1 above and the ** comment below it.

#10 Published: 04/04/00; Revised: 10/27/2002 level for the circuits in Figs. 1 and 2 were set to within 0.1 dB of each other by adjustment of the loss in A1 and A2.

Operation at the LSLCP: Operation at the LSLCP can be said to occur when the detector is operating half way between its linear and square law response mode, the point where the two areas overlap equally. At this point there is a sqrt(2) dB change in output power for every 1.0 dB change in input power.

Operation at power levels below the LSLCP point: As input power levels are lowered, the DDIPL approaches 10*log{[I2/(I2 + Is)]}-6 dB.

Operation at power levels above the LSLCT point: Here, the DDIPL tends to approach zero, but the detector input and output impedance match starts to deteriorate. This is the regime where the mode of detection changes from "averaging" to "peak". (See Article #0, Section 5 for an explanation of this effect.) Re-matching the input and output circuits at these higher input RF power levels recovers the excess loss caused by the mismatch, and results in the performance given by the equations in Article #15A.

Input/Output impedance interaction: When an input signal is present, interaction between the input and output circuit occurs. That is because the attenuation of the attenuators A1 and A2 must become finite and that lets the interaction come through. If the output load R2 is reduced, the input resistance to the detector will be reduced. If the input source resistance R1 is reduced, the output resistance of the detector will be reduced. This interaction is dependent on the strength of the input signal. For greater input signals, there will be less

Four identical diodes in parallel will give a saturation current four times the Is of one alone. For purposes of Crystal Set design, diodes should not be placed in series. SPICE simulation shows that if two identical diodes are connected in series, the combination will perform the same as one of the diodes alone, but having a doubled value for n. This increased value of n will reduce weak signal sensitivity.

In a particular crystal radio set Is can vary quite a bit without a great effect on performance. One can be in error by several times and still get good results. Too high an Is reduces selectivity on weak signals. Too low a value reduces sensitivity to weak signals and causes excessive audio distortion.

Many times the question is asked, "What is the best diode to use?" The answer depends on the specific RF source resistance and audio load impedance of the Crystal Set in question. At low signal levels the RF input resistance and audio output resistance of a detector diode are equal to 25,700,000*n/Is Ohms (current in nA). For minimum detector power loss at very low signal levels with a particular diode, all one has to do is impedance match the RF source resistance to the diode and impedance match the diodes' audio output resistance to the headphones by using an appropriate audio transformer. The lower the Is of the diode, the higher will be the weak signal sensitivity (volume) from the Crystal Set, provided it is properly impedance matched to it's circuit (see article #1). This does not affect strong signal volume. There is one caveat to this, however. It is assumed that the RF tuned circuits and audio transformer losses don't change. This can be hard to accomplish. It is assumed that the Rs, diode junction capacitance, n and reverse leakage are reasonable. If the diode you want to use has a higher Is than the optimum value, tap it down on the tuned circuit. If the diode you want to use has a lower Is than the optimum value, change the tank circuit to one with a higher L and lower C so that the antenna impedance can be transformed to a higher value and repeat step #1.

If you don't have a diode of the proper calculated Is, you can simulate what the result would be if you did have one by doing the following: Put a small voltage in series with the DC load resistor ground return (see point #4 below). If your diode has too low an Is, biasing the diode in the forward direction will improve sensitivity. If your diode has too high an Is, biasing the diode in the reverse direction will improve sensitivity. See Article #9 on the home page on how to build and use a "Diode Detector Bias Box".

.

3. Estimate the audio effective impedance of magnetic phones as 6 times the DC resistance. Alternatively, build the "Headphone Effective Impedance" measuring device described in Article #2 and use it to determine the headphone impedance. Call this impedance Zh.

4. The average audio impedance of the headphones should be transformed up to the value Rr by an appropriate audio transformer. The step-down impedance transformation ratio needed in the transformer is Rr/Zh. When connecting the transformer high impedance winding to the diode, put a parallel RC (a benny) in series with the ground connection . This will insure that the DC load on the diode can be made the same as the audio AC load. A good value for the R should be about equal to Rr. It's best to use a pot so that the value can be optimized at different signal levels. For minimum audio distortion at medium and high signal levels, the DC load on the diode should be the same as the AC audio load. The value of the C should be large enough to fully bypass the R for audio.

is not the case for silicon pn junction or germanium point contact diodes.

A New diode detector equivalent circuit, with a discure law crossover point.

*** Calculations for a RWDDC using equations #6 and *2an given in Article #15A. These equations assume perfect impedance matching at the input and output.

Part 2b: An alternative DDEC.

An alternative 'diode detector equivalent circuit' (DDEC2) can be formed by moving the tank circuit T from its position shown in Fig. 2 to the left hand terminal of diode Di and moving the bypass capacitor C2 to the right hand end of resistor Rs. This equivalent circuit always operates as a peak detector, so no 'excess loss' need be accounted for. The loss for attenuators A1=A2, at any input signal level may be calculated from equations #3n and #6 in Article #15A. Loss for A1=A2=5*log(DIPL from equation #3n) dB. The input impedance (S11) of the DDEC2 approaches that of the RWDD at high and low input power levels. Its input resistance at intermediate power levels is always lower than that of the RWDD.

Part 3: Further Discussion of the Linear-to-Square-law Crossover Point.

The RF input resistances in the simulations of the DDEC (Fig.2) are within 17%, 17% and 8% at the low, medium and high power inputs respectively, of the simulated resistances of the RWDC (Fig 1). The DDIPL values at each input power

Type of analysis	RF Input Voltage V1, in mV peak	DC Output Voltage V2, in mV	DC Output Current I2, in nA	RF Input Power P1, in dBW	DC Output Power P2, in dBW	DDIPL, |S21|, in dB	Sum of the Attenuation, A1+A2, in dB	'Excess loss', above A1+A2
RWDD simu.	1.0126	0.001231	0.0018208	-127.22	-176.50	49.28*	---	---
RWDD simu.	16.165	0.3149	0.4660	-103.16	-128.33	25.17*	---	---
RWdd simu.	64.923	4.849	7.176	-91.08	-104.59	13.54*	---	---
RWDD simu.	258.97	51.34	75.97	-79.06	-84.09	5.03*	---	---
RWDD simu.	4142	1312.4	1942.1	-54.98	-55.97	0.99*	---	---
DDEC simu.	1.0123	0.001231	0.0018209	-127.22	-176.50	49.28	42.36	6.92**
DDEC simu.	16.198	0.3162	0.4679	-103.14	-128.30	25.16	18.67	6.49**
DDEC simu.	64.942	4.867	72.02	-91.08	-104.59	13.54	9.20	4.34**
DDEC simu.	259.12	51.79	76.64	-79.06	-84.01	4.95	2.80	2.15**
DDEC simu.	4143	1309.4	1937.7	-54.98	-55.96	0.98	0.20	0.78**
Calculated values***	1.0120	0.001231	0.0018210	-127.22	-176.49	49.27	---	---
Calculated values***	16.123	0.31500	0.4662	-103.18	-128.33	25.15	---	---
Calculated values***	64.72	4.867	7.176	-91.11	-104.59	13.48	---	---
Calculated values***	257.3	51.22	75.80	-79.32	-84.09	4.77	---	---
Calculated values***	3841	1307.4	1934.7	-55.64	-55.97	0.33	---	---

Table 1

* The n of real world diodes is never 1.0. Actual values of good detector diodes are usually between 1.03 and 1.10. The input and output power values given in the data group for the RWDD can be corrected if n is over 1.0 by adding 10*log (n) dB to the P1 and P2 figures. Keep in mind that n and (Is) are most always independent of current for Schottky diodes. This

A good value is C=5/(pi*2*300*Rr). The parallel RC will have less effect on reducing distortion or affecting selectivity when receiving loud signals if the transformed headphone load on the diode is lower than the diode output resistance, than if it is higher. For info on the impedance transformation ratios of various transformers see Article #5. The audio transformer should have a low insertion loss. Try to obtain one with less than 2 dB loss from 300-3300 Hz when measured at low Crystal Set signal levels. See Article #5 for info on how to measure transformer Insertion Loss.

5. Connect up the new diode and transformer and the parallel RC. Trim up the value of the R in the parallel RC for the least audio distortion on a loud signal. There should be an improvement in low signal volume and high signal audio distortion as well as better selectivity.

#4 Published: 10/02/99; Last revision: 08/22/2002

ARTICLE 05

Low-loss impedance matching for magnetic and piezoelectric headphones, measurements on many audio transformers, and a transformer loss measurement method

Quick Summary: This Article discusses the use of audio transformers with crystal radio sets and gives the results of loss measurements on several of them. A method for measuring insertion power loss is also described.

Many crystal radio set designs provide impedance step down taps on the final RF tuned circuit. If the diode is connected to one of these taps, its loading on the tuned circuit is reduced and selectivity is improved. Too much of a step down also reduces sensitivity. RF tuned circuit loading by the diode is affected by the diodes' Saturation Current, the headphone effective impedance and the signal level. One can reduce the loading effect of headphone effective impedance and of high signal level by transforming the headphone impedance up to a value that matches the audio output resistance of the diode detector itself. This approach can keep the selectivity high and also increase the sensitivity of the crystal radio set. For info on measuring headphone effective impedance see article # 2.

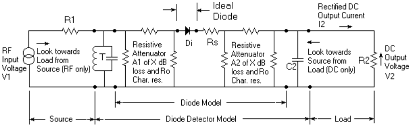
It is important that the diode sees a DC load equal to its AC audio load. This will permit connecting the diode to a higher tap or maybe to the top of the tuned circuit. The result will be to maintain selectivity and reduce audio distortion for medium and especially for strong signals. Diodes of lower Saturation Current can be tapped up higher on the tuned circuit than those of higher Saturation Current and, all else being equal, will give higher receiver sensitivity. See articles #0, #1, #4 and #15.

remain constant no matter what value the attenuation it is set to.

* The source and load resistances of the detector are set equal to the characteristic resistance of the attenuators.

Table 1 shows three groups of data: SPICE simulations of the RWDDC and the DDEC, and a set of calculated values from equations appearing in Article #15A. Data is shown for three input power levels for each data group The levels are: 1) The input power that will operate the RWDDC at its LSLCP [Plsc(i)], 2) 1/128 the value of Plsc(i), and 3) 128 times the value of Plsc(i). These power levels are believed to be correct if the input and output impedances of the detectors are impedance matched, using appropriate values for R1 and R2. Actually, in the simulations, R1=R2=Ro=0.0256789*n/Is. This causes the required input power for the desired output to be somewhat greater than if input and output were perfectly matched. The attenuators A1 and A2 in the DTEC are set equal to each other, and to a value that causes the output power of the DDEC to closely equal that of the RWDDC. SPICE parameters for the diode in the RWDDC and "Calculated values" are: (Is)=38 nA and n=1.0*. The "calculated values" assume impedance matched conditions. The SPICE circuit simulation program "ICAP/4" from Intusoft was used in all simulations.
Data for three different data groups, including loss in attenuators A1 and A2 and the 'excess' loss.

of the RWDD. The input resistance of the DDEC is always higher than that of the REDD. This equivalent circuit seems to work for signals from well below the LSLTP point up to levels just before "Diode Reverse Breakdown Current" comes into the picture.



DDEC to be Simulated in SPICE.
Table of three data groups

Some definitions and conditions that apply to Fig. 2 follow:

* Di is an ideal diode. It has zero forward resistance and infinite reverse resistance. That is, it can pass any amount of current in the forward direction with no voltage drop, and it will conduct no current in the reverse direction, no matter how much voltage is applied. Rs represents the series parasitic resistance of the real world diode (Dr) being modeled. It is shown for completeness, but has negligible effect on the results at the values encountered in crystal radio set operation (5 to 50 Ohms) and will be ignored.

* A1 and A2 are "constant resistance" attenuators of equal attenuation, X dB. Their loss is dependent on the strength of the received signal power. The attenuators each have a characteristic resistance Ro. Is is the saturation current of the real world diode Dr in Fig.1. n is its ideality factor. Note: When a "constant resistance" attenuator is driven by and loaded by a resistance value called its "characteristic resistance", its own input resistance and output resistance

Table of Contents

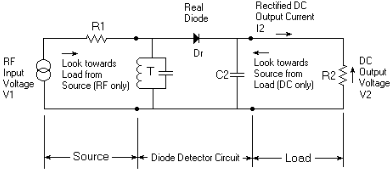
Part 1 - Setup for switchable Transformation Ratios using the A.E.S. P-T157
or equivalent transformers

The sensitivity improvement mentioned above will only be attained if the audio transformation is performed with a low insertion loss audio transformer. For experimental purposes one of the best transformers I have found is the P-T157 from Antique Electronic Supply. What immediately follows is the description of two switchable circuits that can supply various transformation ratios for driving a 12k Ohm load. This is the nominal AC impedance of most 2,000 DC Ohm headphones, as well as many piezoelectric ceramic earpieces. Later on, specific non-switched configurations are shown for several different transformers. Since this Article was written, A.E.S. has stopped selling the P-T157. Results close to those shown below can be obtained using the A.E.S. P-T156, Stancor A-53 or most any 3:1 turns ratio tube-type inter-stage audio coupling transformer. A good description of this type of transformer is: A transformer designed for plate-to-grid inter-stage coupling, having a 3:1 turns ratio, and specified for a 90k to 10k ohm impedance transformation. Henceforth, in this Article, this type of transformer will be referred to as a "3:1 AIT".

Note that the switched transformation ratios shown below vary by a factor of about four from one to another. Note also that an impedance mismatch of 2:1 gives an insertion loss of only 0.5 dB. This means that all values of diode output resistance from 12k Ohms up to 750k can be utilized, with a mismatch insertion loss of no more than a maximum value of 0.5 dB, plus the transformer loss. Measured transformer loss is about 1.0 +/- 0.5 dB from 300- 3300 Hz at the 63 times ratio and about 0.5 +/- 0.2 dB at the 16 and 4 ratios. Note: The transformation ratio on the H switch position is shown as 63 instead of 72 because of shunt resistive losses in the transformers. On this switch position the diode sees the 12k headphones transformed to 750k, not 860k. T1 and T2 are

will see an RF resistance (at fo) equal to the junction resistance of the diode at zero bias. At this very low signal condition the detector input resistance is not affected by any changes made to R2. The value of this junction resistance is the slope of the diode V/I curve at the origin. From a differentiation of the ideal diode equation, the numerical value of this resistance is: (0.0256789*n)/Is ohms at a temperature of 25 degrees C. Let's call this Ro. Is and n are parameters in the ideal diode equation. (For a discussion of Is, n, etc., see the text after the schematics in Part #1 of Article #1). From the load resistance R2, looking back toward the input, one sees the same resistance value Ro, and it is independent of any changes at the source. Now look at Fig. 2. Here, the real world diode has been changed to a theoretical ideal diode and two attenuators, A1 and A2, of characteristic resistance Ro have been are added. If V1 becomes zero, the attenuators A1 and A2 must be set to infinite attenuation to enable the model to duplicate the behavior of the circuit in Fig.1. When an input signal is applied, the values of A1 and A2 must become finite. The DDIPL is equal to the sum of the loss of each attenuator plus the impedance interface loss between the ideal diode Di and each attenuator, as well as any mismatch loss between R1 and the detector as well as between R2 and the detector (See ** after Table 2). SPICE simulation shows that the diode detector equivalent circuit does a pretty good job modeling the operation of a real world diode detector. To verify this, one can perform a SPICE simulation of Fig.1 and Fig. 2 with V1, R1 and R2 the same in each case. The attenuation value of A1=A2 dB must be set to a value that causes the output, V2, in Fig. 2 to be the same as in Fig.1. The input impedance match of the two simulations differ from each other by less than 14% over an input power range of 48 dB, centered at the Linear-Square-Law Transition (LSLCP) point. This is the main area where the results from the DDEC simulation differ from those how weak or strong the input signal. See Article #8 for an illustration of typical waveforms. A reactance value for the tank capacitor equal to less than one hundredth of the value of R1 will be sufficient. The DC resistance of the tank inductor should be small enough so that no appreciable DC voltage will appear across it. A value less than one hundredth of the value of R2 will be sufficiently small. This assures that all of the output DC power goes into R2. The bypass capacitor C2 has a very low reactance compared to the load resistor R2 at the frequency fo. Since C2 acts as a short circuit across R2 at the frequency fo, all of the RF voltage across T will appear across the Diode. The time constant, R2*C2 should be long compared to the time for one cycle of fo.

Part 2a: Discussion of the new Diode Detector Equivalent Circuit.



RWDDC to be Simulated in Spice.
Diode detector to be simulated in SPICE

To gain an understanding of the Diode Detector Equivalent Circuit (DDEC), first consider the following line of thought: See Fig. 1. Let the input RF voltage V1 become very low. V1, at a frequency fo, looking toward the load resistance R2,

preferably Antique Electronic Supply P-T157 transformers. Alternatively, one can use most any generic "3:1 AIT". One will get a small amount more loss with the alternatives, mainly at 300 Hz and when the signal is weak. C2 can be used to peak up response at 300 Hz if a generic "3:1 AIT" transformer is used. C2 can be omitted if the P-T157 transformers are used. Experiment with values around 0.02 uF. Sw1 and Sw2 are DPDT slide or toggle switches. R can be a 1 Meg pot. It is used to set the diode DC load resistance to be equal the transformed AC load impedance.* A log taper is preferred. Set R for the lowest audio distortion and best selectivity on strong signals. The diode load at DC must be the same as for AC audio signals for best results. This setting has little effect with weak signals, however. C1 should be about 0.05 uF. See the later part of Article #1 for info on determining transformer winding polarity and how to reduce the effect of inter-winding capacitance.
* The first time anyone has suggested placing a parallel RC in series with the diode to enable adjusting its DC load resistance equal its average AC load may have been in Article #1. Some people call it a "benny".

Schematic showing a switcheable two transformer array using PT-157's.

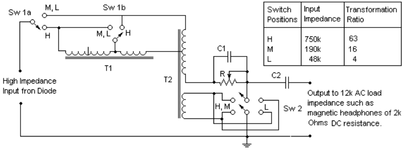


Fig. 1

There is no need to transform headphone effective impedance up to as high as 750k unless the RF tuned circuit, when loaded with the antenna, has a resonant resistance of around 750k Ohms. It is very difficult to attain an impedance this high. The diode also would have to have the appropriate Saturation Current of about 38 nA. For experimental purposes, if a transformed impedance of no higher than 380k is desired, a one transformer circuit should be used as shown below. This will prove more practical in real world applications. R may be a 250k or 500k Ohm pot, preferably with a log taper. The transformer insertion loss remains below 1.0 dB from 0.3-3.3 kHz with output loads between from 6k to 24k Ohms when using the A.E.S. P-T157. Keep in mind that the saturation Current (Is) of the diode should be such that the diode's (Weak signal) RF input resistance is about equal to the (Antenna loaded) RF tuned circuit resonant resistance and also to the transformed headphone effective impedance. This diode resistance is equal to (0.0257*n)/Is. Is is in Amps. For more information on this, see article #4 listed on the home page.
Schematic showing one Switcheable PT-157 Transformer.

A real world diode is a two terminal device. The "real world diode detector circuit" will be modeled as a "two port, four terminal device" having a pair of terminals for the input and another for the output. One of the input terminals is the "hot" input terminal; the other is "low". One output terminal is the "hot" one, the other is "low". The two "low" terminals are connected together and usually to ground. Please note, that in the topology of the two schematics shown below, the "Diode Detector Circuit" and the "Diode Detector Equivalent Circuit" both include the tank T and the bypass capacitor C2 as an integral part of the detector. Also, look at the circuits in this way: The tank circuit, looking towards the output, sees the diode as a one-end-grounded shunt load since the output bypass cap is a short at RF. The output load resistor, looking back towards the input, sees the diode as a one-end-grounded shunt DC resistive source since the input side of the diode is shorted to ground by the tank.

See Fig 1. The detector tank circuit T is modeled as lossless and resonant to the input frequency fo. Losses in a real world tank can be accounted for by using Thevenin's Theorem to calculate the appropriate changes in V1 and R1. This leaves the circuit topology unchanged. See Article #1 for more on this subject. The value of the tuning capacitor in T is sufficiently large so that essentially no harmonics of fo can appear across T. This assures that the "pendulum-like resonator effect" of a high Q circuit will be available to supply the narrow, high-current pulses the diode requires every cycle when strong signals are handled. Another advantage is that tank-voltage-waveform peak clipping by diode conduction is essentially prevented when the current pulses are drawn. All this assures that the input impedance to the detector will be linear over one cycle of RF and the input current to, and voltage across the tank T will always be sinusoidal, no matter valid ways one can use in thinking about how a circuit works, the better becomes one's understanding of that circuit.

This analysis applies to an AM detector fed by a CW RF sine wave voltage of frequency fo: It has a peak (not RMS) value equal to V1 and an internal source resistance of R1. The "maximum available RF input power" is called P1 (see section 2 in Article #0 for info on "maximum available power"). The DC output power delivered to the load resistor R2 is called P2. The DDIPL (in dB) is equal to ten times the log of the ratio between the two powers P2 and P1.

This approach can also be used to model how a diode detector behaves with an AM modulated input signal by performing a SPICE simulation three times. Once with the RF signal equal to the value of the desired modulated wave's envelope minimum value, once with the signal equal to the carrier value and once with the signal equal to the crest value. The three DC output voltages give the minimum, carrier equivalent and peak value of the demodulated output audio wave.

***** Please do not skip this next paragraph! *****

To understand the new diode detector equivalent circuit, one must abandon the usual way of thinking the about the diode in a detector. Instead, one must think about the "diode detector circuit". This circuit includes a tank circuit T, the output capacitor C, as well as the diode. The shunt input reactance of the circuit is assumed to be zero at all frequencies except fo, the frequency to which the tank is tuned. The input resistance at fo will be discussed later. The output reactance of the circuit is assumed to be zero at all RF frequencies. The output resistance will be discussed later.

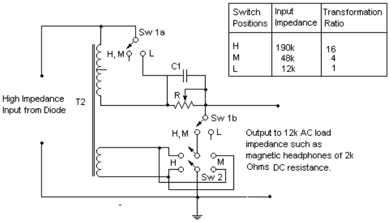


Fig. 2

Part 2 - Practical Fixed Transformation Ratio Setups using the A.E.S. P-T157, PT-156,
Stancor A-53 or similar Audio inter-stage Transformer "3:1 AIT"

The following schematics show various connections for the transformers mentioned above. The connections are arranged to provide various diode audio frequency load impedances from headphone loads of either 12,000 or 1,200 Ohms AC impedance. The 12,000 Ohm connection is appropriate for most magnetic headphones of 2,000 Ohms DC resistance and many piezo earpieces. The 1,200 Ohm connection is used when driving a series connected set of typical sound powered elements.

As stated before, it is important that the diode have a DC load of the same value as its average AC load. This can easily be accomplished by placing the parallel combination of a pot with

an audio bypass capacitor (a "benny") in series with the lead marked 'RC' between the points marked "x--x" (see below). The pot should have an audio taper and be connected as a rheostat. 0.5 to 1 Meg is usually a good value. The value of the capacitor depends upon the impedance reflected into the transformer primary from the headphones. A value of 0.05 uF or more is usually OK. The pot should be adjusted to minimize distortion and improve selectivity when receiving strong signals. Its setting has no effect when receiving weak signals.
Various connections of transformers to enable different impedance transformations

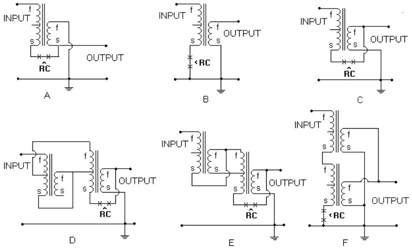


Fig. 3

First, some help. In schematics A-F it is important to properly phase the windings. For best performance at the high end of the audio band, one should minimize the effect of transformer inter-winding capacitance. This is most important to do when using circuits D or E, but has little effect when using circuits A, B, C, or F. To do this, the start and finish leads of the

ARTICLE 10

A New Diode Detector Equivalent Circuit, with a Discussion of the Linear-to-Square-Law Crossover Point: the signal level at which the detector is functioning midway between linear and square-law operation

Quick Summary: The purpose of this article is to describe and and then compare a new diode detector equivalent circuit (DDEC) to a real world detector circuit (RWDDC), such as might be used in a crystal radio set. This equivalent circuit uses an ideal diode. Comparisons are made using SPICE simulations of the two circuits. Calculations using equations given in Article #15A are also supplied for comparison. The concept of the Linear-to-square-law crossover point (LSLCP) in the relation between output DC and input AC power is introduced (not to be confused with the exponential relationship of DC current to DC voltage in a diode).

Part 1: General Description of a Diode Detector.

The new diode detector equivalent circuit (DTEC) is based on the idea that a detector diode imbedded in a proper circuit can be thought of as a 'black box' device that converts RF power into DC power. Some power is lost in the process and that is called Diode detector insertion power loss (DDIPL). This approach completely avoids such concepts as duty cycle, pulse current, bypass capacitor charging and non-linear instantaneous voltage/current relationships. It is also consistent with the material given in Article #1. The peak-detector, capacitor-charging-current line of thought is good when signal levels are high enough to assure that true peak detection occurs. It is not very useful when signal levels are low. However, when all is said and done, the more different

What to do now? If the optimum diode has a higher Is than your present one, several identical diodes can be paralleled to create the equivalent of one of higher Is. For instance, five in parallel will have an Is five times that of one alone. If you have several different diodes, experiment with them. Maybe you can find one that does not need a bias for best results. In recent years many different types of diodes called 1N34A have been sold. Their Is values vary all over the lot.

Some final comments: Using the Bias Box to reduce the effective Is of a diode that has a high Is does not work very well if a large reduction is needed. The reason is that diodes of high Is naturally have higher back leakage and a lower reverse breakdown voltage. This causes losses and audio distortion when the RF voltage across the diode swings to reverse polarity every RF cycle. Less sensitivity and selectivity is the result. When one increases the effective Is of a diode that has a low Is by applying a forward voltage bias, this problem does not occur. Some other things that will cause some optimized diodes to work worse than others are: High series resistance (Rs), high diode barrier capacitance (this reduces high frequency performance compared to that at lower frequencies) and high reverse leakage current.

#9 Published: 03/30/00; Last revision:01/24/07

transformer coils must be properly connected. In the schematics shown above, the start and finish of the transformer windings are indicated by "s" and "f". The start of the low impedance winding of a PT-156 or P-T157 transformer is the blue lead. The finish is the red lead. The green lead next to the red lead is the start of the center tapped high impedance winding. The green lead next to the blue lead is the finish. If Stancor A53-C transformers are (is) being used, the color coding is different. The start of the low impedance winding is the red lead, the finish is the blue lead. The green lead of the center tapped winding next to the blue lead is the start, and the green lead next to the red lead is the finish.

The insertion loss values shown below are were measured using A.E.S. P-T157 transformers. Stancor A53-C or generic "3:1 AIT" units will perform somewhat worse, as mentioned above. If you are going to use a generic "3:1 AIT", keep in mind that all or most of the extra insertion loss at 0.3 kHz can be eliminated by using the correct capacitor in series between the transformer and headphones.

Table 1 - Insertion Loss for Various Impedance Transformations when Driving Magnetic Headphones (2k ohms DC resistance) or Piezoelectric Earpieces of about 12k Ohms AC Impedance. For A.E.S. P-T157 and generic "3:1 AIT" transformers . Frequency Range is from 0.3-3.3 kHz.

SOURCE IMPEDANCE RANGE	LOAD IMPEDANCE	CIRCUIT	INSERTION LOSS
25k-70k Ohms	12,000 Ohms	A	0.6-1.0 dB
70k-150k	12,000	B	0.3-0.8
150k-250k	12,000	C	0.2-0.6
250k-500k	12,000	D	0.3-1.2
500k-700k	12,000	E	0.5-1.5

A very low loss transformer that can be used to transform a 1 Megohm source down to closely impedance match a 12k AC ohm load is the small UTC O-15 'Ouncer' transformer. Its insertion power loss is less than 1 dB from 0.3-3.3 kHz.

Table 2 - Insertion Loss for Various Impedance Transformations Driving Sound Powered Headphones of 1.2k Ohms AC impedance. For A.E.S. P-T157 and generic "3:1 AIT"s. Frequency Range is from 0.3-3.3 kHz.

SOURCE IMPEDANCE RANGE	LOAD IMPEDANCE	CIRCUIT	INSERTION LOSS
16k-54k Ohms	1,200 Ohms	C	1.1-1.3 dB
43k-130k	1,200	F	1.3-1.9

Here are general specifications for the A.E.S. P-T157, Stancor A-53C and generic "3:1 AIT" Inter-stage transformers: Single Plate (10,000 Ohms) to push-pull grids (90,000 Ohms). Overall turns ratio: 1 to 3 Primary to Secondary. Max. Primary DC: 10 mA. These transformers are still relatively cheap and usually available at Hamfests, personal junk boxes and Used Component Vendors.

Part 3 - Transformer configurations for use mainly with Sound Powered Headphones

Now we will talk about some other transformer configurations that are suited for use with Sound Powered phones: UTC LS-10, UTC A-10, UTC A-12, Amertran 923A and UTC C-2080, as well as many others. The UTC A-10 and A-12 have the same terminal impedance specifications as the LS-10 and will probably perform similarly. Some of these transformers are currently quite expensive. For some lower cost options, see Part 5 of this article for some generic transformer specs., or consider the last two connections shown in the chart above. Shown below are loss measurements using a physically very

transformer primary, if one is used. If no transformer is used, connect the terminals of T1 to the crystal radio set ground and the cold end of the headphone headset. Also make sure that the connection where the Bias Box is inserted is well bypassed for RF and audio.

To operate the box, snap the switch to OFF, disconnect the Hot T1 connection from the crystal radio set and adjust the DIODE DC LOAD pot to the DC load desired (See articles #1 & 4 or pick 100k Ohms to get started). The DC load resistance can be measured across the terminal strip labeled T1 when its hot lead is disconnected from the crystal radio set. Reconnect the Hot T1 connection to the crystal radio set DC return. Tune in the weakest station you can copy. Snap the switch to ON. Adjust the BIAS pot for the the greatest volume. Tune in the strongest station you can get. Adjust the DIODE DC LOAD pot for the least audio distortion. Disconnect the antenna. Connect a DVM to terminal strip T2. If you find a voltage there, that is an indication that your diode is not optimum. A diode having a different Is could work the same, but without the need for any bias. If your diode is biased in the forward direction, the optimum diode would have a higher Is than your present diode. A reverse bias indicates that the optimum diode would have a lower Is. As stated before, the relationship between the required bias (Vbias), the Saturation Current of the original diode (Isor) and the Saturation Current for the optimum diode (Isop) is: Vbias = 0.026*n*ln(Isop/Isor) volts. Some illustrations: To change Is by five times, the bias Voltage required is about 0.044 Volts. To change it by 25 times, the Voltage is about 0.088 Volts. Note: When adjusting the BIAS pot from the optimum position, moving toward forward bias reduces volume, sensitivity and selectivity. Moving toward reverse bias increases selectivity, reduces volume and sensitivity and adds audio distortion.

higher will be the very weak signal sensitivity. At low signal levels, the Id*Rs expression is small and can be neglected.

To change the detector performance of a diode of Is = Isor (original) to the performance of a diode of Is = Isop (optimum), a DC bias voltage must be inserted in series with the diode. The required bias voltage is: Vbias = 0.0257*n*[ln (Isop/Isor)]. ln represents the natural logarithm of the expression following it. This equation is accurate if the values of Is and n do not change as a function of diode current. This assumption is correct for the Schottky diodes I have checked. Some germanium ones I have checked do not accurately follow the Shockley equation. They tend to have high values of Is such as 500 nA or more. Germaniums having Is values in the 100-200 nA range do seem to follow the Schockley equation well. At high currents, Is increases from its value at low currents. The Vbias equation is given for information only and is not used in the following experimental procedures. Whether a Schottky, germanium or other diode is used, a convenient way to 'tune' the Is of a diode is to use the "Diode Bias Box". It effectively enables one to change a diodes' effective Is (and therefore its operating impedance) by merely turning a knob on a box. The Diode Bias Box also enables one to determine the best diode DC and AC load impedance.

Here is an interesting relationship that applies to most Schottky diodes: A Schottky diode detector having a saturation current of (Is1) that has no external DC current bled into it will perform, as a diode detector, identically to that of another diode having a saturation current of (Is2) if a DC current (Ib) equal to (Is1-Is2) is bled into it.

To use the Bias Box, connect the terminals labeled T1 to the crystal radio set ground and the cold end of the audio

small, but very low cost transformer, the MOUSER TM-117 as well as two excellent small low loss transformers from the CALRAD line.

At the end of this Section (#3), measurements on a combination of two transformers are be shown that enable 900k to 1200 ohm and 470k to 1200 ohm impedance transformation.

Six measurements on the TM-117 are shown. The first test of the transformer is with the input and output resistance values specified by the Manufacturer, but at a low output signal level. The second is for a TM-117 driven and loaded by the resistances of 24k ohms primary and 300 ohms secondary instead of 50k ohms primary and 1k ohms secondary. The 24k ohm level is close to that delivered by a generic 1N34A diode when detecting a weak signal. The next three measurements are for four TM-117 transformers interconnected to give a transformation ratio four times greater than one gets from one transformer alone. The primaries are connected in series and the secondaries are connected in series/parallel. The resultant primary and secondary are connected as an autotransformer. Results are given from measurements made at three output power levels. The last measurement is with the transformers connected for a 1,200 Ohm output instead for a 300 Ohm output. Most Sound Powered elements I have seen have an AC impedance of about 600 Ohms when averaged over the frequency range of 0.3-3.3 kHz. When used as a 1,200 Ohm transformer load, the two elements should be connected in series. When used as a 300 Ohm load, the elements should be connected in parallel. Remember that the insertion loss near 0.3 kHz can usually be reduced by placing a proper capacitor in series with the connection from the transformer to the sound powered headphones.

Table 3 - Insertion Loss Values for Various Transformers driving Sound Powered Headphones or elements of 300, 600 or 1,200 Ohms effective Impedance. Frequency Range is from 0.3 - 3.3 kHz.

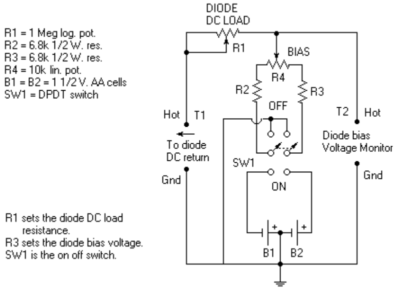
Transformer Model #	Source Impedance (Primary) in Ohms	Load Impedance (Secondary) in Ohms	Connections	Insertion Loss Range in dB
UTC LS-10	120,000	300	G	0.4
UTC LS-10	270,000	300	H	0.7-1.0
UTC LS-10	270,000	1,200	I	0.7-1.0
UTC LS-10	430,000	1,200	G	0.8-1.6
UTC C-2080*	330,000	300	J	0.8-1.2
UTC C-2080*	540,000	600	J	1.3-1.9
UTC C-2080*	820,000	1200	J	2.1-3.2
AMERTRAN 923A	680,000	300	K	1.0-1.8
AMERTRAN 923A	680,000	1200	L	1.0-1.8

* The UTC-2080 is rated by the manufacturer for transforming between source/loads of 100 and 100k Ohms. A similar transformer made by the Stanley Company (TF-1A-10-YY) is (or was) offered by the Fair Radio Sales Co. as #T3/AM20, for about $7.95. They are recommended as good all around choices for driving 300 ohm sound-powered phones (SP elements in parallel) in good quality crystal radio sets.

Table 4 - AC parameter values of the UTC 2080 and Stanley TF-1A-10-YY transformers, measured at the 100 ohm terminals (#1 & 2).

Transformer name	Magnetizing Inductance	Resistance at resonance	Resonant frequency	Distributed capacitance	Inter-winding capacitance
UTC 2080	395 mH	2.25k ohms	0.98 kHz	69.6 nF	39 pF
Stanley TF-1A-10-YY	595	3.15k	1.65	15.7	66

To obtain the approximate magnetizing inductance, resonant resistance and distributed capacitance for the UTC and Stanley units that appears at terminals 3 and 4 (other winding open-



Schematic of the Diode Detector Bias Box.

A detector diode having particular saturation current (Is) can be biased to perform almost exactly the same as a diode having a different Is. This statement assumes that the diode conforms to the classic Shockley diode voltage/current relationship: Id = Is*{exp[(Vd-Id*Rs)/(0.0257*n)]-1)}, at room temperature. Id is the diode current in Amps. Is is the diode Saturation Current. "exp" means: raise the base of the natural logarithms (2.718...) to the power of the expression following. Vd is the voltage applied to the diode in volts. Rs is the fixed series parasitic resistance of the diode in ohms. n is the Diode Ideality Factor (Emission coefficient) and is dimensionless. It is usually between 1.05 and 1.2. The lower the value of n, the

ARTICLE 09

Build the Crystal Set Diode Detector Bias Box: a simple and easy way to determine if one's diode is optimum for weak signal reception, or it should have a higher or lower axis-crossing resistance (0.026*n/Is ohms)

Quick Summary: The 'Diode Detector Bias Box' enables one to check whether the diode being used in a crystal radio set has optimum characteristics for that set. The optimum detector will deliver the greatest low-signal sensitivity.

A detector diode, in order to deliver the highest sensitivity and lowest audio distortion, must be properly impedance matched to its RF source. It must also be matched to the correct (for that diode) audio and DC load resistances. See Articles # 0, 1, 5 and 15a for more info on this subject. How can one know for sure that the diode used in one's own crystal radio set is the best one for it? Another way of putting it is: Does my diode have the Saturation Current (Isopt) that the optimum diode, for my set, would have? An easy way to find the answer is to build and use the diode Detector Bias Box.

Picture of the Diode Detector Bias Box.

circuited), multiply the values of the first two parameters above by 1000 and divide the distributed capacitance by 1000. This is because of the 1:1000 impedance transformation ratio. Note: Terminals 1 and 2 are marked as the "100" ohm terminals, 3 and 4 are marked "100k". Bear in mind that the magnetizing inductance of these transformers can vary appreciably from sample to sample because of the low (or no) air-gap design used in the laminations.

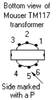
Table 5 - Insertion Loss of MOUSER TM-117 Transformer(s) using various
Interconnections and Load Resistances

Transformer Model #	Source Impedance in Ohms	Load Impedance in Ohms	Connections	Output Power Level	Insertion Loss in dB: 0.3*, 1.0, 3.3 kHz
Mouser TM-117	50,000	1000	M	-60 dBm	11.1, 1.8, 5.7
Mouser TM-117	24,000	300	M	-48 dBm	4.9, 1.2, 2.6
4 Mouser TM-117	100,000	300	N	-72 dBm	5.3, 1.5, 4.6
4 Mouser TM-117	100,000	300	N	-42 dBm	4.9, 1.1, 4.6
4 Mouser TM-117	100,000	300	N	-12 dBm	1.7, 1.7, 4.6
4 Mouser TM-117	100,000	1,200	O	-52 dBm	4.7, 1.2, 4.3

*Some or all of the loss at 0.3 kHz can be eliminated by coupling the transformer to the headphone load through a series capacitor. This makes a high pass filter with a cutoff frequency at or somewhat below 0.3 kHz out of the components, instead of having a just a plain old shunt parallel RL 6 dB/octave roll-off response. The components of the filter are the shunt inductance of the transformer, the series capacitor and the shunt inductance of the headphone impedance. A value around 2 uF is usually good if the headphone effective impedance is 300 Ohms (elements connected in parallel). A value around 0.5 uF is good if the headphone effective impedance is 1,200 Ohms. (Elements connected in series) One must experiment with different values because the inductance and effective impedance of different elements varies from Mfg. to Mfg. Of course, this principle may be employed at other impedance levels such as the 12k ohms in a Brandes Superior headset, when used with an appropriate transformer (see the paragraph above Fig. 1).

Table 6 - Terminal Connections for UTC, AMERTRAN and MOUSER TM-117 Transformers

G. Join 1 & 3, 4 & 6, 8 & 9. Input is 7. Output is 1. Ground is 4 and 10.
H. Join 2 & 3, 4 & 5. 8 & 9. Input is 7. Output is 2. Ground is 4 and 10.
I. Join 3 & 4. 8 & 9. Input is 7. Output is 2. Ground is 5 and 10.
J. Input is 3. Output is 2. Ground is 1 and 4.
K. Join 1 & 3, 2 & 4, 6 & 7. Input is 8. Output is 1. Ground is 4 and 5.
L. Join 2 & 3, 6 & 7. Input is 8. Output is 1. Ground is 4 and 5.
M. Input is 4. Output is 1. Ground is 3 and 6
N. Take four TM-117s and label them W, X, Y and Z. They will be connected in an autotransformer configuration. Join W6 to X4, X6 to Y4, Y6 to Z4. Join W1 to Y1. Join X3 to Z3, Join W3 to X1. Join Y3 to Z1. Connect a parallel RC from Z6 to W1. Input is W4. Output is Y1. Ground for input and output is Z3. For an explanation of why to use the RC, see the second paragraph after the first graph in article #1.
O. Take four TM-117s and label them W, X, Y and Z. They will be connected in an autotransformer configuration. Join W6 to X4, X6 to Y4, Y6 to Z4. Join W3 to X1, X3 to Y1, Y3 to Z1, Connect a parallel RC from Z6 to W1. Input is W4. Output is W1. Ground for input and output is Z3. For an explanation of why to use the RC, see the second paragraph after the first graph in article #1. Desirable but optional: Connect X1 to the center of the 1,200 Ohm load (junction of two 600 Ohm sound-powered elements connected in series). This eliminates a narrow spurious 1 dB dip in the frequency response at about 1.2 kHz.



The transformer loss figures for the UTC and AMERTRAN transformers were measured at an output power of about -60 dBm. Performance is retained at output power levels much less than -60 dBm. A voice signal at this power level will be quite soft, but understandable through most sound powered headphones.

The MOUSER transformer deserves special discussion since it is so low in cost (available at Mouser Electronics (http://www.mouser.com). Frequency response and distortion: The loss figures at two different power levels for a

If we were to restore an optimum impedance matched condition by adjusting the input source resistance to 495k ohms and the output load resistance to 990k ohms (by changing the input and output impedance transformation ratios), the power loss would be further reduced to 0.22 dB. See Part 5 of Article #0 for further discussion on the subject of input and output resistance of diode detectors operated in their peak-detection mode.

#8 Published: 02/13/00; Last revision: 11/25/01

calculates as: ((4.08)^2)/700k = -46.23 dBW. Detector power loss goes down to: 46.23 - 45.54 = 0.69 dB. The input resistance is now even lower than before.

The green output voltage at test point Y3x is 4.08 volts, kind of low compared with the 6.25 volts we would get with a perfect input impedance match. Why is this? The input and output resistances of a diode detector both approach the value (0.026*n)/Is Ohms at low signal power levels. (n and Is are diode parameters used in SPICE. n is called the diode Ideality Factor, or Emission Coefficient. Is is called diode saturation current.) Is is defined as the current that is asymptotically approached in the diode back bias direction before extraneous leakage factors or reverse breakdown comes into play. It also has a major effect at an on the amount of current a diode will pass in the forward direction at any specific applied Voltage.

As we have seen, as signal input power increases, the quality of the RF impedance match starts to degrade. The AC input resistance to the diode detector decreases from the value obtained in the first well matched low power level simulation. Interestingly, the output resistance increases. 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 very low power loss. The rule here is that the DC input resistance of an ideal diode peak detector is one-half the value of the output load resistance. Also, the output resistance is equal to two times the value of the input source resistance. Further, since in this example the detector now approaches being a true peak detector, the DC output voltage approaches the square root of 2 times the value of the RMS input voltage. This relationship is necessary in an ideal peak detector so that the AC input power can equal the DC output power with no power lost in the diode (No free lunch).

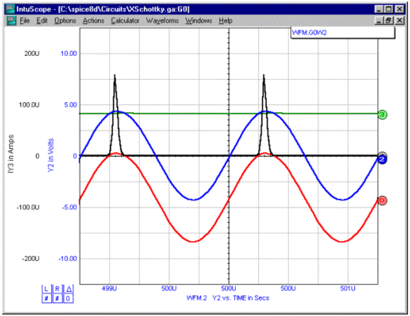
TM117 purchased in March of '00 are as follows: Output power level of +15 dBm: 2.8 dB @ 0.3 kHz, 1.9 dB @ 1.0 kHz and 6.4 dB @ 3.3 kHz. Output power level of -60 dBm: 11.1 dB @ 0.3 kHz, 1.8 dB @ 1.0 kHz, 5.7 dB @ 3.3 kHz and 5.4 dB @ 0.6 kHz. The 0.3 kHz loss is greater at a power level of -60 dBm than at +15 dBm. Why? The core laminations of the TM-117 (and many other very small transformers) have low permeability at the low magnetic flux levels generated by the -60 dBm signal. This low permeability is called initial permeability. The initial permeability, in combination with other factors, results in the transformer having a specific shunt inductance (at low signal levels). This shunt inductance controls the low frequency roll-off of the transformer. At higher flux levels (signal levels), but before saturation occurs, the permeability increases to an "effective permeability" value which can be several times greater than the initial permeability. This means that the transformer shunt inductance is higher at the higher signal level and the low frequency roll-off is much reduced. There may be some production unit-to-unit variation in the low frequency response of the TM117. One that I bought about a year ago showed 2.5 dB less loss at 0.3 kHz than the one tested above. Some low frequency harmonic distortion is generated in the changeover region from initial to effective permeability. This can easily be seen on a 'scope, especially at 300 Hz sine wave. I doubt that it would be very noticeable in actual crystal radio set use.

One can see from lines three, four and five of data in the TM-117 Insertion Loss Chart above, that the loss at 0.3 kHz, relative to that at 1.0 kHz, gets less as the output power level is increased. The loss at 1.0 kHz is minimum at the -42 dBm output power level. The greater 1.0 kHz loss at the -72 dBm power level is caused by the reduced shunt inductance as explained above. The increase in 1.0 kHz loss at -42 dBm occurs because the core is getting closer to saturation. The loss at 3.3 kHz in the four-transformer configuration is greater than that for one transformer shown on line 2 because the primary-to-secondary capacitance of transformers A and B is effectively connected from high impedance points and ground, thus rolling off the high end response. The single transformer in line 2 is wired so that the primary-to-secondary capacitance is not in shunt across the primary to ground.

The CALRAD line of small transformers offers two types that are suitable for use in transforming a high diode detector output resistance down to 300 or 1200 Ohms to drive SP phones. Their insertion loss is quite low and within a fraction of a dB of that of the UTC LS-10. One distributor of CALRAD transformers is Ocean State Electronics, 6 Industrial Drive, P.O. Box 1458, Westerly, RI. http://www.oselectronics.com/ (they call these transformers (Mini Audio Transformers). The two transformers are #45-700, spec'd to transform 100k to 1000 Ohms and #45-703, spec'd to transform 200k to 1000 Ohms. They sell for about $5.95 ea. The following chart shows the measured performance of a single transformer and of combinations of two. Lines #1 and 2: Primaries are in series, secondaries in parallel. Line #3: Primaries are in parallel, secondaries in series. Performance is very good, especially so, considering the price.

Table 7 - Insertion Loss of certain CALRAD transformer(s) as single units, and with two connected together.

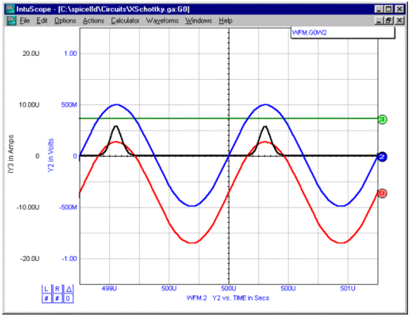
voltage at test point Y3x is 361 millivolts. Forward current is now drawn over about 24% of one cycle time. The input available power, as stated
Before, is -65.54 dBW. The output power is ((0.361)^2)/700k = -67.30 dBW. Detector power loss is: 67.30 - 65.54 = 1.76 dB.



Current and Voltage Waveforms at a High Input Power Level.

Now it looks as if we are getting much closer to peak detection. The peak positive voltage applied to the diode at test point Y2 is 4.30 volts. The detected DC voltage at test point Y3x is 4.08 volts (only about 5% less than the 4.30 volt peak). The diode forward conducts only during about 12% of the cycle time of the 1.0 MHz wave. As stated before, the input available power is -45.54 dBW. The output power

Finally, note that there is no peak detection going on. The diode output voltage, measured at test point Y3x is only 15.7 millivolts even though the peak forward voltage applied to the detector is 61.9 millivolts. Input power as stated above is -85.54 dBW. The output power is ((0.0157)^2)/700k = -94.53 dBW. Insertion loss = 94.53 - 85.54 = 8.99 dB.



Current and Voltage Waveforms at a Moderate Input Power level (linear/square-law breakpoint).

Here, the input voltage at test point Y1 is 1.25 volts, but the voltage across the LC tank circuit, as measured at test point Y2 is only 494 millivolts, not 625 which would be the case if we had a perfect impedance match. This shows that the detector input resistance is now lower than 700k Ohms. Diode operation is getting closer to peak detection. The green output

Line #	Transformer Model # (s)	Source Impedance in Ohms	Load Impedance in Ohms	Output Power Level in dBW	Insertion Loss in dB: 0.3*, 1.0, 3.3 kHz
1	Two 45-700	110k	300	-54	1.2, 0.9, 0.9
2	Two 45-703	270k	300	-54	1.9, 0.9, 1.3
3	Two 45-703	51k	1200	-54	1.6, 1.0, 1.0
4	One 45-700	91k	1200	-54	1.8, 1.0, 0.9
5	One 45-703	220k	1200	-54	3.7, 1.4, 1.3
6	Two 45-700	350k	1200	-54	3.0, 1.1, 1.8
7	One 45-703+ one 45-700	510k	1200	-54	4.8, 1.5, 2.5

* See asterisk just below the preceding Mouser transformer table.

Note: The hot lead of the high impedance winding should always be the red lead. The hot lead of the low impedance winding should be the white lead. The high impedance windings are connected in series in lines 1, 2, 6 and 7. They are in parallel in line 3.

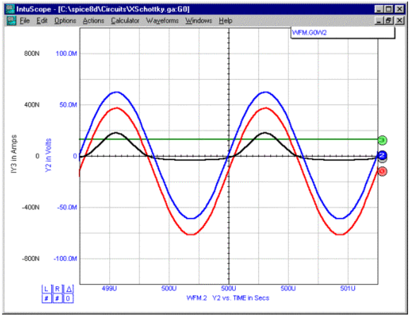
The low impedance windings are connected in parallel in lines 1, 2, 6 and 7 with leads of like color connected together. The hot low impedance output connection is to the white leads. The other two joined leads go to ground. The low impedance windings are connected in series in line 3.

One must properly phase the windings when two transformers are used. The lead from the hot end of the high impedance winding of transformer #1 should be connected to the diode. Its cold lead should go to the hot lead on transformer #2. The cold lead of transformer #2 should go to a parallel RC, the other end of which should either go to ground in lines 1, 2, 3, and 7 or to the hot low impedance output in line 6 (autotransformer connection). In line 7, transformer #1 should be the 45-703.

A UTC O-15 'Ouncer' transformer can be combined with a Bogen T-725** to make an excellent low loss transformer assembly for matching a wide range of headphone impedances, as well as an 8 ohm speaker, to a 1.35 Meg source resistance. See Fig. 4a. Insertion Power Loss is 2.6 dB @ 0.3 kHz*, 1.2 dB at 1 kHz and 1.7 dB @ 3.3 kHz. Note, that for these measurements, the housing of the O-15 was left ungrounded to eliminate the approximately 20 pF stray capacity from terminal #4 to the case. This reduces loss at 3.3 kHz, but might introduce hum in some applications. An alternative connection that matches to a 1 Meg source is shown in Fig.4b. This lowers the transformation ratio to reduce the effect of external stray capacitance-to-ground from crystal set components connected to the high impedance point. It also reduces sensitivity to hum pickup if the case of the O-15 is left ungrounded. The result is a small 0.5 dB loss reduction loss at 3.3 kHz when that stray capacitance is 16 pF. The Bogen T-725 transformer is available from Dave Schmarder's "Hobby Electronic Parts" at: http://www.1n34a.com/catalog/parts.htm . The UTC Ouncer O-15 transformer is hard to find, but sometimes shows up on e-Bay.
*See the asterisk at the end of the Mouser insertion loss table above.

The R1 C1 combination is sometimes called a "benny". It is used to reduce audio distortion sometimes encountered on strong stations. A good value for the pot, R1, is 1-3 Megs, preferable with an audio taper. C1 is not critical. A value of 0.1 uF is suggested.

A typical diode for use with this transformer assembly is a one having a saturation current of about 22 nA, such as the Agilent 5082-2835 or HSMS-2820. The weak signal audio output



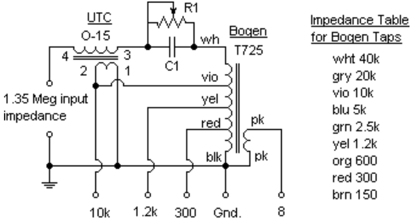
Current and Voltage Waveforms at a relatively Low Input Power Level.

The black curve shows the diode current. The other three curves all use the same scale on the vertical axis. The blue curve shows the voltage at the test point Y2. This is the voltage across the tuned circuit. It has a peak value of 61.9 millivolts, about 1/2 that at test point V1. This shows that the detector has an input resistance of about 700k Ohms. There is a good input impedance match here. The red curve shows the voltage across the diode. Note that where it is positive, a forward diode current flows for about 42% of the time for one cycle of the 1.0 MHz wave. Note that where it is negative, a reverse diode current flows. This reverse current flattens out and if a higher input signal was used, it would flatten out at about 38 nanoAmps, the saturation current of the diode.

and n are not constant and do vary with diode current. Measurements made on one 1N34A shows Is and n values of 2.7E-6 and 1.64 at 320 uA which drop to 1.21E-6 and 1.34 at 32 uA, then down to 6.6E-7 and 1.05 at 1.8 uA. Schottky diodes seem to have constant values for n and Is.

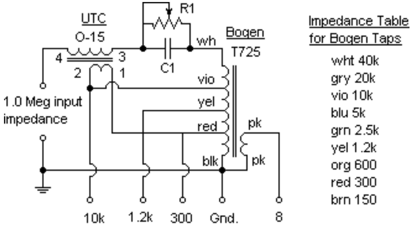
The SPICE netlist above shows, as the input, a 1.0 MHz sine wave of peak amplitude 0.125 volts for V1. (This Input signal level is 2.74 dB less one that would operate the detector half way between the linear and square law modes. At this lower input signal power level the insertion power loss of the detector is 7.12 dB). The first of the three simulations will be done with an input sine wave of 0.125 volts peak for V1 as shown in the netlist. The second simulation will use a 1.25-volt peak sine wave. The third will use a 12.5-volt peak sine wave. The respective available input powers are: -85.54 dBW, -65.54 dBW and -45.54 dBW (dB below one Watt).

resistance of such a diode, as well as the input audio resistance of this transformer assembly shown in Fig. 4a are each about 1.35 Megs, making their parallel combination about 675k ohms. If the total shunt capacitance at this point is about 70 pF, audio frequencies above 3.3kHz will be attenuated by more than 3 dB. This capacitance consists of the sum of the windings capacitance of the transformer assembly referred to its input, wiring capacity to the diode and the junction capacity of the diode, etc, all in parallel with the RF bypass capacitor of the detector (C9 in Fig. 5 in Article #26). When using this transformer assembly, if the audio high frequencies seem deficient, try reducing that capacitor (C9), or maybe eliminating it and depending upon the other capacitive elements for RF bypassing. If one possesses good high frequency hearing, a higher impedance tap than normal may strengthen the highs. This is because the impedance of magnetic headphones is not constant. It rises as frequency increases, therefore, a better impedance match will exist for high frequencies when a higher impedance tap than normal is used. See "The effect of source impedance on tone quality" in Article #2.



Some Output Impedance Values in Ohms

Fig. 4a



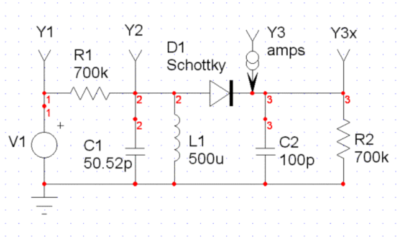
Some Output Impedance Values in Ohms

Fig. 4b

**Note re the Impedance Table for the Bogen T725 transformer, above: The Bogen Speaker Matching transformer is designed to be used when one wishes to drive multiple 8-


```
C:\spice8d\Circuits\XSchottky.cir Setup1
*#save V(1) V(2) @R1[i] @R1[p] @C1[i] @L1[i] V(3)
@D1[id]
*#save @D1[p] @C2[i] @R2[i] @R2[p]
*#viewtran iy3
*#alias iy3 @d1[id]
*#alias y1v(1)
*#viewtran y1
*#alias y2v(2)
*#viewtran y2
*#alias y3xv(3)
*#viewtran y3x
.TRAN 31.25n 502u
*#save all
.OPTIONS reltol=0.00001
.OPTIONS vscale=0.25
.PRINTTRAN IY3
.PRINTTRAN Y1
.PRINTTRAN Y2
.PRINTTRAN Y3x
V1 1 0 SIN 0 0.125 1meg 0 0 0
R1 1 2 700k
C1 2 0 50.52p
L1 2 0 500u
D1 2 3 _HP2835
.MODEL _HP2835 D BV=15 CJO=0 EG=0.69 IBV=2.5e-5
IS=38nA
+ N=1.03 RS=6.4 VJ=0.56
C2 3 0 100p
R2 3 0 700k
END
```

Note 1: In regard to the accuracy of the SPICE diode model, some diodes, notably the 1N34A are unusual. The values of Is signal input power level. Two modes of operation for a detector have been defined: Linear and square law. Linear operation is said to occur when a change of input power (in dB) causes an equal change in output power. Square law operation is said to occur when a given small change in input power (in dB) causes double that change in output power. Where is the breakpoint between linear and square law operation? SPICE simulation gives the answer, to the extent that SPICE and the diode models are accurate (See Note 1. after SPICE netlist). An input power sufficient to cause the rectified DC current to equal to the saturation current (Is) of the diode is an indication of operation half way between linear and square law. The detector power loss at this level is 7.1 dB.



Schematic of Diode Detector Circuit using SPICE.

The Intusoft ISpice netlist shown below is automatically generated by the SpiceNet program after the schematic and parts values are entered into the program.

ohm speakers, distributed over a wide area, from a single audio source, e.g., a public-address system. In order to limit I-R losses in the distribution line, the system is operated at a higher than normal impedance and voltage. For example, a 100-watt amplifier, with a 49-ohm output impedance, will deliver 70 volts to the line. At each speaker, a transformer is used to connect the 70-volt line to the speaker voice-coil. The line-to-voice-coil transformer has multiple primary taps to allow the power delivered to each speaker to be individually adjusted to suit system requirements. The taps on such transformers are often labeled with the power they will draw from a 70-volt, or sometimes a 25 volt line. When adding an 8 ohm speaker to a system with a Bogen T725, it is first connected to the pink wires of the T725. One then selects a tap to connect to the audio source, depending upon how loud one wants the speaker to be (taps white through brown), with black as common. The step-down turns ratio between [(white through red)-to-black] and [pink-to-pink] sets the volume. Basically, to accomplish its objective, the T725 was designed to transform an 8 ohm speaker load to one of various other impedances between 150 to 40k ohms as shown in the Table in Fig 4, above.

Now let us get out of the context of sound distribution and into crystal set audio impedance matching. Consider the tapped black through white winding of the Bogen T-725 as just a conventional autotransformer designed for maximum efficiency (low loss) when used at the impedances shown. There is no "magic" in the "nominal" values of 150 through 40k ohms assigned to the various colored taps. The autotransformer can just as well be used at impedances a multiple higher or lower than those shown, although with some extra insertion power loss. It seems that the convention of assigning impedance values to the taps has fostered some confusion. For instance, one should not think that the "right" impedance to connect to the green wire is 2.5k ohms. In the BT- Ultimatch (see Part 4, below), the settings of the input and output switches connected to T725 should be set for minimum insertion power loss. Probably the 'nominal" impedance of the output tap used will be close to that of the load but not necessarily the same.

Part 4 - The BT-UltiMatch, a modified version of Steve Bringhurst's UltiMatch. Insertion power loss and input resistance measurements when using a selection of various 'Stanley-type' transformers or an UTC O-15 for T1 for the input transformer are displayed

Steve Bringhurst's UltiMatch (see http://www.crystalradio.net/soundpowered/matching/index.shtml#UltiMatch) on Darryl Boyd's Site provides a convenient low-loss way to provide audio impedance matching between the output resistance of a diode detector and the average impedance of headphones or a speaker. The BT-UltiMatch is somewhat different from Steve's, the differences being the replacement of Steve's SW5 with a single-pole 10 point rotary switch and provision for switching in an UTC O-15 in place of the "Stanley-type" transformer used for T1. The two10-point rotary switches enable one to connect any of the taps of T2 to the secondary terminals of either T1 or T3, as well as to the output capacitors C2-C6. This provides for a greater range of impedance transformation and a further reduction in power loss (especially at high source resistances) than when one is limited to using only the brown or red wires of T2. Provision is also made for switching in a UTC O-15 Ouncer transformer in place of the Stanley-type unit.

across the tank circuit. (Appreciably all to the tank circuit voltage, therefore, appears across the diode).

6. The output load resistance may seem to be a high value for headphones. It is assumed that in practice, the headphone impedance will be transformed up to that value by a low loss audio transformer. It is also assumed that the transformer primary has an appropriate capacitor bypassed resistor in series with it. The purpose of this is to insure that the audio load on the diode has the same DC as AC value.

7. The RF and AF load resistances used in the simulation will seem quite high. This is because the average unloaded shunt resistance of the loop in my single tuned loop receiver is 700k Ohms, and I am using it in the simulation that follows.

8. The diode junction capacitance is set to zero in the netlist. This has no effect on the operation of the detector if C1 is retuned to take account of this fact. Experimentation is now more convenient since a change of C2 will have no effect on tuning.

9. The diode parameters are specified so as to produce an RF input resistance of 700k Ohms when operated in a detector circuit and driven by a low available power source of, say, -85 dBW.

A basic crystal radio set diode detector schematic is shown below. An Intusoft SPICE simulator will be used in three separate simulations to measure circuit currents and voltages. The calculations from the simulations will show that the detector insertion loss approaches zero at high input power levels and that it goes up sharply as the input power goes down below a certain point. This loss will be minimized if the input and output resistances of the detector are impedance matched. The following discussion assumes that the RF source and both the DC and Audio AC load are matched to the diode at a low voltage squared divided by resistance. In the impedance matched condition, because of the 2 to 1 voltage division from the source resistance and load resistance, one-half of the internal voltage V will appear across the load resistance. The actual power absorbed by the load will be, as indicated in the preceding relation: P = ((V/2)^2)/R = (V^2)/(4R). Half of the power delivered to the series combination of the source resistance and the load resistance will be delivered to the load. The other half is dissipated and lost in the source resistance. In the crystal radio set case the input voltage is AC RF voltage. If the input voltage is referred to by its peak value (Vp) as it is in SPICE, 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 P = (Vp^2)/(8R). This is the equation that will be used to calculate available input power to the detector, from the source.

Here are some definitions, assumptions and explanations:

1. The internal resistance of the antenna is transformed up to the equivalent parallel resistance R that is used in the simulation. The tuned circuitry used to do this is not shown.
2. The single tuned circuit used is assumed to have an infinite Q. A finite Q will cause an increase in insertion loss.
3. "Diode Detector Power Loss" is defined as the ratio of DC output power dissipated in the output load resistance to the RF input "available power". (Expressed in dB)
4. The L/C ratio of the tuned circuit L1, C1 is sufficiently low so that no appreciable harmonic voltages will be developed across it by the detection action of the diode.
5. The RF bypass capacitor C2 is sufficiently large so that the RF ripple voltage across it is small compared to the voltage

Two sets of measurements were made using 300 and 1200 ohm resistive loads (typical average impedance of SP headphones having both elements connected in parallel is 300 ohms, in series, 1200). The input voltage at 1 kHz was series-connected through a selection of source resistors to the input of the BT-UltiMatch. The level of this voltage was adjusted to produce about 10 mV RMS across the output load of 1200 ohms (5 mV when measuring with a 300 ohm load). See Part 5 of this Article for info on how the BT-UltiMatch loss measurements were made.
Schematic of BT-UltiMatc

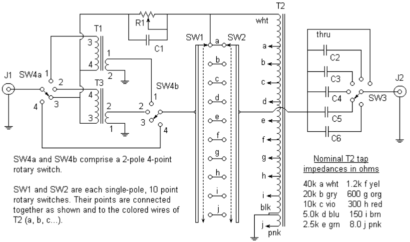


Fig. 5 The BT-UltiMatch, a modification of Steve Bringhurst's UltiMatch, a low-loss variable audio impedance matching device.

Parts list

* T1 - "Stanley-type" 100k-100 ohm transformer, available from Fair Radio Sales as # T3/AM-20
* T3 - UTC Ouncer O-15 1 meg-10k ohm transformer
* R1 - 2 or 3 Meg log taper pot or, a 1 Meg log taper pot with provision for switching in a series resistor to increase the

total resistance to 2 Megs if low saturation current diodes (having a high axis-crossing resistance) are to be used

* C1 - 0.22 uF capacitor
* C2 - 4.7 uF non-polar capacitor
* C3 - 2.2 uF non-polar capacitor
* C4- -1.0 uF non-polar capacitor
* C5 - 0.47 uF non polar capacitor
* C6 - 0.22 uF non polar capacitor
* SW1, SW2 - Single pole 12 position (only 10 used) rotary switches. Designate SW1 and SW2 switch positions with the nominal impedance values given in the Table in Fig 5
* SW3 - One pole, 6 position rotary switch
* SW4 - Two pole, 4 position rotary switch

Operation of the BT-UltiMatch

Four modes of operation are provided by SW4:

1.

This most common mode uses a "Stanley type" transformer as the input transformer, T1. For use with a wide range of diodes having moderate axis-crossing resistances, such as the ITT FO-215 germanium diode (Is~100 nA).

2.

Uses UTC O-15 Ouncer transformer for T1. Use with diodes having high axis-crossing resistances such as the Agilent 5082-5235 Schottky diode (measured Is~15 nA, not the spec sheet value of 22 nA), or 2-3 in parallel.

3.

Does not use input transformer T1. Use with diodes having low axis-crossing resistances such as the 1N34A germanium diode (Is~600 nA).

4.

ARTICLE 08

Crystal radio diode detector power loss with current and voltage waveforms, as determined from a SPICE Simulation

Quick summary: This Article shows diode detector voltage and current waveforms and how they change as a function of signal strength.

In this article I am going to show an analysis of the operation of a crystal radio set detector using a SPICE simulator. The detector voltage and current waveforms will be shown for three different input "available power" sources. These sources will supply either -85.54, -65.54 or -45.54 dBW (number of dB's below one Watt) power to a matched load. Each power source is made up of a pure voltage source combined with a resistance. (The combo could also be referred to as a "voltage source with an internal resistance"). In each case the available input power, the output power and detector insertion loss will be shown. Conformance to or deviation from the usually assumed peak-detector model will be investigated. The change in input resistance with change in input power will also be examined.

Here is a derivation one needs to know in order to understand the rest of this article. The concept of "available power": If one has a voltage source V with an internal resistance R, then the load resistance to which the maximum amount of power (Pa) can be delivered is itself equal to R. Pa will be called the "maximum available power". Any load resistance other than one equal to the source resistance R will absorb less power from the source. This applies whether the voltage is DC or AC (RMS). An equation for power absorbed in a resistance is

V/I Graph of 1N34A Diode. V/I Graph of 1N34A Diode.
V/I Graph of 1N34A Diode. V/I Graph of 1N34A Diode.
Forward Conduction Graphs of 1N34A and 1N914 Diodes.
Reverse Conduction Graphs of 1N34A and 1N914 Diodes.

#7 Published: 01/08/00; Last revision: 05/25/00

Does not use input transformer T1. Use for general purpose impedance transformation, working with T2 only (SW1 and the "benny" - R1,C1 are out of the circuit).

J1 should be connected with a very short cable (when using very low Is diodes) to the diode output of a crystal set, the usual audio source. J2 is connected to headphones or a low impedance speaker. The "nominal" impedance designations on SW1 and SW2 are for reference and do not always represent the settings for the least insertion power loss. When starting out to match real world headphones or a speaker to the diode output of a crystal set, do the following:

1. Select a setting for SW4, depending upon the estimated Is of the diode being used (see Tables 1 and 2 in Article #27 for some help on this).
2. Set the "nominal impedance" of SW2 to the expected average impedance of the output load (see Article #2 for info on how to obtain this info).
3. Set SW3 to the "thru" position.
4. Adjust SW1 for loudest volume.
5. Adjust R1 for minimum distortion.
6. The optimum settings for SW1 and SW2 are interactive. Try higher and lower settings for SW1 , then tweak SW2 to see if greater volume is available, then iterate.
7. Try different settings for SW3 to see if bass response improves.

Table 8 - Insertion power loss in a BT-UltiMatch @ 1 kHz using a selected Stanley transformer TF-1A-10-YY "D" or an UTC Ouncer O-15 transformer for T1. Measurements are made with different series-connected source resistances and with load resistances of 300 or 1.2k ohms. SW1 and SW2 were adjusted for maximum output

Transformer	Source and load resistances in ohms, insertion power loss in dB				
Stanley "D"	2.2 Meg, 300; -2.59	1.0 Meg, 300; -1.41	470k, 300; -0.70	220k, 300; -0.69	100k#, 300; -0.94
UTC O-15	2.2 Meg, 300; -1.61	-	-	-	-
Stanley "D"	2.2 Meg, 1.2k; -2.50	1.0 Meg, 1.2k, -1.37	470k, 1.2k; -0.94	220k, 1.2k; -0.95	100k#, 1.2k; -0.84
UTC O-15	2.2 Meg, 1.2k; 1.54	-	-	-	-

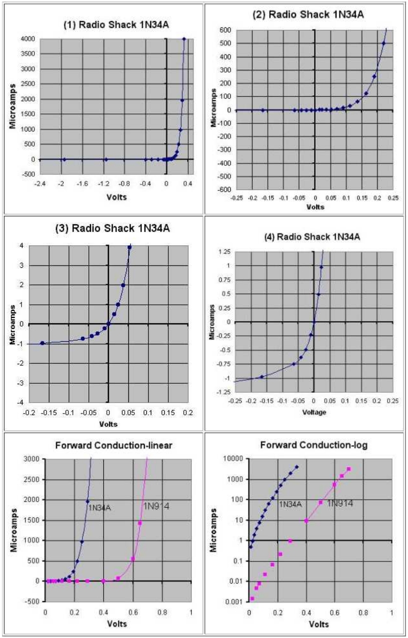
Table 8A - Nominal** settings for SW1 and SW2 taps in ohms for minimum insertion power loss, using a UTC or selected Stanley transformer for T1

Transformer	Source resistance, load resistance of 300 and [T1 and T2 nominal tap settings] in ohms				
Stanley "D"	2.2 Meg; 600, 300	1.0 Meg; 300, 300	470k; 1.2k, 1.2k	220k; 600, 1.2k	100k#; 40k, 150
UTC O-15	2.2 Meg;, 10k, 150	-	-	-	-
Transformer	**Source resistance, load resistance of 1.2k and [T1 and T2 nominal tap settings] in ohms**				
Stanley "D"	2.2 Meg; 1.2k, 1.2k	1.0 Meg; 600,1.2k	470k; 300, 1.2k	220k; 300, 2.4k,	100k#; 40k, 600
UTC O-15	2.2 Meg; 10k, 600	-	-	-	-

See Table of Impedance taps in Fig 5 showing nominal impedances for the various taps on the Bogen T-725 transformer

SW4 set to position 3

Table 9 - Insertion power loss (dB) in a BT-UltiMatch @ 1 kHz using various transformers for the input transformer with various series-connected source resistances. Load resistance of 1.2k ohms. SW1 and SW2 were adjusted for maximum output



For RF diode detectors to work, one needs a device that has a non-linear V/I curve. In other words, the slope of the V/I curve must change as a function of applied Voltage. The slope must be steeper (or shallower) at higher voltages and shallower (or steeper) at lower voltages than at the quiescent operating point. To clarify this, look at curve #3. As a low-level signal detector, this diode will rectify if biased at -0.025, 0.0 or +0.025 volts. The difference is that the diode resistance at the -0.025 Volt operating point is higher than that at 0.0 Volts. It is lower at +0.025 than at 0.0 Volts. If one places a straightedge on the screen of one's PC monitor, tangent to the curve at -0.025, 0, and then +0.025 Volts, one can measure a slope of about 80k Ohms at -0.25 Volts, 40k Ohms at 0.0 Volts, and 20k ohms at +0.25 Volts. The rate-of-change of slope as a function of voltage (second derivative) is less at -0.025 Volts than at +0.025 Volts. This means that the detection sensitivity when biased at -0.025 Volts will be less that when biased at +0.025 Volts, even it the input and output are properly impedance matched.

Transformers	Series-connected source resistance in ohms; insertion power loss in dB					
Stanley "C"	-	1.5 Meg: -1.95	1.0 Meg: -1.51	470k: -0.98	220k: -0.97	-
Stanley "Z"	-	1.5 Meg: -2.26	1.0 Meg: -1.75	470k: -1.03	220k: -1.00	-
Stanley "D"	2.2 Meg: -2.50	1.5 Meg: -1.81	1.0 Meg: -1.37	470k: -0.94	220k: -0.95	100k#, -0.84
UTC C-2080 "A"*	-	1.5 Meg: -1.73	1.0 Meg: -1.36	470k: -0.99	220k: -0.96	-
UTC C-2080 "B"	-	1.5 Meg: -1.90	1.0 Meg: -1.46	470k: -0.93	220k: -0.96	-
No name TA-18.071*	-	1.5 Meg: -1.81	1.0 Meg: -1.42	470k: -0.92	220k: -0.97	-
UTC O-15 Ouncer Tx	2.2 Meg: -1.54	-	-	-	-	-

* The polarity of one winding in each of these transformers was reversed during its manufacture. Measurements were made with terminals 1 and 2 interchanged to correct the condition. The polarity assumption used in the BT-UltiMatch for a Stanley-type T1 is as follows: If an AC voltage is applied from terminals 3 to 4, a voltage of the same polarity will appear from terminals 1 to 2. It is recommended that anyone building a BT-UltiMatch check the polarity of the internal terminal connections of the T1 transformer with a 'scope.

Table 10 - Input resistance of a BT-UltiMatch driven from various series-connected source resistances, terminated by a 1200 ohm load and adjusted for minimum loss

Transformer used for T1	Series-connected source resistance	BT-UltiMatch input resistance
Stanley TF-1A-!0-YY "D"	100k	171k#
Stanley TF-1A-!0-YY "D"	220k	286k
Stanley TF-1A-!0-YY "D"	470k	517k
Stanley TF-1A-!0-YY "D"	1 Meg	751k
Stanley TF-1A-!0-YY "D"	2.2 Meg	1.10 Meg
UTC O-15 Ouncer	2.2 Meg	1.68 Meg

SW4 set to position 3

Notice, since the transformers are mostly not used at their design-center impedances, settings for minimum insertion power loss do not always coincide with an impedance matched condition.

Part 5 - How to Measure the approximate Insertion Power Loss of any Audio Transformer
or Compare its Performance to that of an Ideal no-loss Transformer

The equipment needed are an audio sine wave generator, an assortment of resistors (preferably not a resistance box), and a high sensitivity scope or DVM. The use of a vertically calibrated scope is preferable to a DVM, as one can see that the waveform is clean and without appreciable hum or noise. A difficulty with this approach is that one must make sure that the scope decade attenuator as well as the 10X switch on the probe are accurate. I use the scope probe switched to 1X when reading the low voltage secondary voltage and to 10X setting when measuring at the higher voltage primary. The high input impedance of the probe prevents excessive loading of the high impedance primary, thus reducing the voltage there and causing an incorrect reading. When a DVM and a 'scope of adequate sensitivity are available, the best approach, and the one I now use, is to connect them in parallel. This provides the relatively high precision of a reading with the DVM along with the ability to monitor the voltages for purity (low distortion of the sine wave measuring wave form and low noise).

Connect the hot lead of the generator to the high impedance primary of the transformer through a resistor of value equal to Rs. Rs should be equal to the expected output resistance of the diode detector. Connect a load resistor of value Zh (expected effective impedance of the headphones) to the secondary. Connect all grounds to a common point.

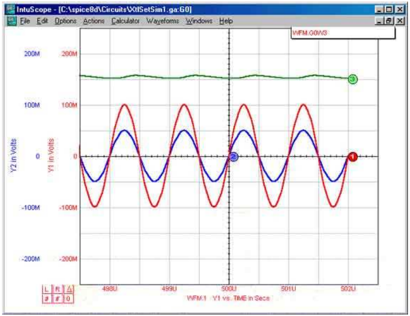
Tune the generator to the first frequency of measurement, say 1000 Hz. Connect the scope or DVM to the low impedance secondary. Adjust the audio generator to as low a level as

ARTICLE 07

Diode Voltage/Current Curves: Does a Specific "Knee" Voltage really Exist?

I think that there is validity to the notion of the existence of a diode "knee" when clamping circuits are considered, and maybe with other circuits. I do not think that there is any validity to the notion of a "knee Voltage" in the forward conduction portion of a diode curve when low signal level detection is considered. The reason is that the apparent Voltage of the knee is an artifact of the Current Scale used in plotting the diode V/I curve. In fact, the shapes of the forward conduction curves of all normal diodes are quite similar, and with no "knee", if the Current scale is logarithmic, not linear. To illustrate this, take a look at the charts below. The first four use a linear scale for the Current axis. The full-scale Current values are: 4000 uA, 600uA, 4uA and 1.25uA. The 1N34A diode is one purchased at Radio Shack with measured Is =2.57uA, n=1.6, and Rs=6.55 ohms. The values of Is and n were calculated from measurements made at an effective diode current of 320 uA. The fifth chart shows the 1N34A and a 1N914 using a linear current scale. The sixth chart shows the two diodes using a log Current scale. The 1N914 has n1.85, Is=2.3nA and Rs=6.0 Ohms.

Graph #1 seems to show a knee at about 0.2+ Volts. A knee at about 0.2 Volts seems a little ambiguous in graph #2. In graph #3 the knee has vanished. Graph #4 seems to show a knee on the current scale in the reverse bias region! The fifth and sixth graphs show a comparison of the 1N34A and 1N914 in the forward conduction region with a linear and then a log Current scale. Note the apparent knees on the linear plot and the total absence of any knee on the logarithmic plot.



The last 4 1/2 Cycles of the Spice Simulation.

#6 Published: 01/02/00; Last revision: 11/23/01

possible while still being able to get an accurate reading of the voltage without error from hum and noise. Read this voltage and call it E3. Now connect the scope probe to the hot end of the primary. Read that voltage and call it E2. Connect the scope probe or meter to the actual generator output (not the transformer hot lead). Read this voltage and call it E1. Calculate insertion loss: Loss=10*log{4*RS*[(E3/E1)^2]/Rl} dB. Also take measurements at 300 and 3300 Hz. If the 300 Hz loss is much greater than the 1000 Hz loss, a transformer with a higher primary inductance is needed. If the 3300 Hz loss is much higher than the 1000 Hz loss, the transformer has too high a winding capacitance for the primary source resistance (RS) selected. If the loss at 1000 Hz is above about 2 dB, a better transformer probably exists. Hopefully all readings will be better than -2 dB.

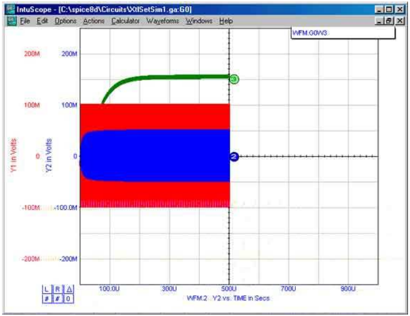
If the transformer is doing a good job of impedance matching RS to Rl, E2 will be about 1/2 the value of E1 and the transformer insertion loss will be at about its minimum. If E2 is much lower than 1/2 of E1, a greater impedance transformation (turns ratio squared) is needed. If the transformer has taps on the secondary, using a lower impedance tap might improve results. If E2 is higher than 1/2 of E1, the impedance transformation ratio is too large and a higher impedance secondary tap should be tried (if available). It is assumed here that reactive mismatch from transformer shunt inductance and distributed capacitance is negligible. It's usually best to make the 1/2 voltage measurement at the frequency of minimum loss (usually about 1 kHz for audio transformers.

In this series of Articles the statement is often made or implied that power loss in an audio transformer is at a minimum when the input is impedance matched. This is not strictly true. If a transformer has internal resistive power loss and is matched at its input, the output will, in general, be mismatched. A simultaneous matched condition at both input and output is usually impossible unless the transformer has no internal losses, or its series and shunt losses are in the correct proportion. A real world transformer delivers its minimum loss when the input and output mismatches (S parameter return losses) are equal. Verification of this condition is both difficult and unnecessary because the two loss values (matched input vs equal mismatch at input and output) normally differ very little. This being the case, one can say, for practical purposes, that the minimum insertion power loss occurs when the input is impedance matched.

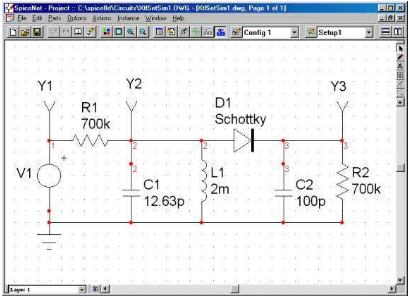
An easy way to compare the performance (loss) of a particular transformer with that of an ideal no-loss transformer of just the right transformation ratio is to build and use the 'Unilateral Ideal Transformer Simulator' described in Article #14.

Tip: If hum and noise are a problem, place the scope (or DVM), signal generator transformer and all leads on a metal ground plane connected to the scope ground. Ordinary kitchen aluminum foil is suitable for the ground plane.

Note: If one has some sort of audio impedance measuring device and desires to measure the shunt inductance of an audio transformer, make sure the measurement frequency is low enough so that the transformer winding capacitance does not interfere with accuracy. A transformer, tested above, that is very good for many crystal radio sets when driving 300 ohm headphones is sold by Fair Radio Sales Co. http://www.fairradio.com/ as #T3/AM20 (similar to UTC #2080). Its winding capacitance is at approximate resonance with its shunt inductance at 1 kHz. This is good design



Input and Output Waveforms from the full SPICE Simulation.



Schematic of Detector Circuit in SPICE.

practice since it minimizes insertion power loss at the (approximate) geometric center of the audio band. A measurement of its unloaded (high impedance winding) Z at 1 kHz yields a result of "infinite" shunt inductance in parallel with a resistance of more than a Megohm. A measurement at 300 Hz gives a result of several hundred Henrys. A measurement at 3300 Hz would show a capacitive, not inductive impedance.

Part 6 - Some practical suggestions on where to get and how to identify transformers
that may perform well with sound powered headphones

Here are some generic transformer specifications, which when met, probably indicate that the transformer will exhibit low insertion loss when used to drive sound powered phones in a crystal radio set. A transformer obtained at a Hamfest, junk box or Surplus Dealer that meets these specs. will probably cost substantially less that the UTC and Amertran transformers. Fair Radio Sales Co. at http://www.fairradio.com/often has suitable transformers available at reasonable prices.

* Wide frequency response specification such as +/- 1 dB, 20-20,000 Hz: A transformer having a Manufacturer's specified operating frequency range from, say, 200-5,000 Hz will probably have several dB more loss than a wide band unit when both are sourced and loaded with resistances that reduce their bandwidth to covering only the 0.3-3.3 kHz range. The reason for this is that we usually want to operate the transformer with primary source and secondary load resistances several times higher than that which the manufacturer specifies. This always narrows the transformer

pass-band. We should shoot for a final pass-band of 0.3 - 3.3 kHz, or so.

* High impedance winding specification: Single grid, preferably push pull grids, single plate or preferably push pull plates. The impedance level, if specified, will usually be between 20,000 and 80,000 Ohms. The high impedance winding is the one to connect to the diode detector output.

* Low impedance winding specification: Low impedance mike, pickup, multiple-line or simply a number between 100 and 1000 (Ohms). Several taps may be supplied to enable various impedance levels. The specification might be: 50, 125/150, 200/250, and 333, 500/600 Ohms. This winding is the one to which the sound powered phones are to be connected.

* The correct low impedance tap to use for connecting the sound powered phones may be calculated as follows:

Decide the audio load resistance to be presented to the detector. Let's select 200,000 Ohms. (See Articles #1 and #4 for info on how to determine this value). Assume that the sound powered elements are connected in series. This will typically result a headphone average impedance of 1,200 Ohms. Calculate the needed impedance transformation ratio as: 200,000/1,200 = 167. Note the Manufacturer's specification for the high impedance winding of the transformer. (If you don't know what it is, estimate 80,000 Ohms.), and divide it by 167. Select the Manufacturer's low impedance winding tap specification (if you have that info) that most closely equals the value calculated above. If you are using the 80,000 Ohm estimate, the desired tap impedance would be 80,000/167 = 479 Ohms. Call this number A. Now check what the result would be if the sound powered elements were connected in parallel. They will now present an effective impedance of 300 Ohms and require an impedance

netlists: XtlSetSim1.cir and XtlSetSim2.cir can be edited and used in any other SPICE simulator.

Keep the following things in mind:

1. Simulations are only as good as the SPICE simulator, its device models and the circuit topology used. The Shockley diode equation agrees well with the results from the Schottky diodes I have checked, even at low currents. With the one germanium 1N34A I have checked, the Shockley diode equation equation works well above about 40 micro-amps but not very well below that. (for a specific voltage, the equation specifies a higher current than the diode delivers.)
2. The tuned circuit L1|C1 must be tuned to resonance. When this is the case, the voltage across the tuned circuit will be in phase with the source voltage V1. If the Intusoft simulator is used, the probe point for V1 is Y1 and the probe point for the tuned circuit voltage is Y2. If one views Y1 and Y2 on the same graph one can check the relative phases of the two voltages. Y3 gives the output voltage.
3. The carrier ripple shown at the output has negligible effect on the average output. A larger value for the filter bypass C2 can reduce the ripple, but at the expense of rise time of the output voltage.
4. When considering the practical application of simulation ideas, keep in mind the advice in articles # 1, 4 & 5 on this Website, especially as regards audio impedance matching.
5. Shown below are the schematic diagram from the schematic editor Spicenet, the simulation of all 4002 cycles of the 1.0 MHz signal at Y1, Y2 and Y3, and the last four cycles of the 1.0 MHz signal. Note that this simulation uses a Schottky diode, not a 1N34A diode.

be about 18dB less or -86dBm. Of course, the impedance values used here are quite high, but they are the values I achieve in my loop crystal radio set. To find out where the -18 dB came from, see Article #1, end of part 1, from the home page.

In XtlSetSim2 the input sine wave is set to a peak value of 0.045 volts. Since the source resistance is set to 16,000 ohms, the power incident on the detector is -47dBm. The output power delivered to R2 is -68dBm, the same as in the XtlSetSim1 example. The audio load used is 16k ohms. Note that the insertion loss in XtlSetSim1 is 68-57=11 dB. The loss in XtlSetSim2 is 68-47=21 dB. XtlSetSim1 requires 10 dB less input than XtlSetSim2 for the same output! XtlSetSim1 uses a diode with a saturation current Is=40 nA and n=1.08. XtlSetSim2 uses a diode of Is=2600 and n=1.6. The second graph in article #1 on this site predicts that the loss difference would be 8dB. This experiment illustrates that a detector using a of a lower Is, if it is matched at the input and output, will have a lower loss than one using a diode of a higher Is. I have found, since this article was written, that in a 1N34A germanium diode, the values of Is and n change at low currents. Is may go down as much as 5 times and n may drop 25% from the values used in simulation XtlSetSim2. (Those values were obtained at an unrealistically high diode current of about 320 uA.) This was not expected. The result is that the germanium diode is unfortunately shown incorrectly and in a very unfavorable light (for crystal radio set use). The simulation in XtlSetSim2 probably should have used an Is of about 700 nA and n should have been about 1.15.

I used the SPICE program from Intusoft called ICAP4WINDOWS demo version. It can be downloaded for free from their Website at http://www.intusoft.com. The

transformation ratio of 200,000/300=667. The desired Manufacturer's tap marking will now be 80,000/667 = 120 Ohms. Call this number B. Pick the number A or B, whichever is closest to an available transformer tap marking. Connect the phone elements appropriately. Note that we are using the transformer at a higher impedance level than that for which it was designed. What we lose is by doing this is audio bandwidth and a small increase of insertion loss. We don't need the 20-20,000 Hz range anyway, do we? What we gain is an ability to transform headphone impedance to a higher value than if we used the manufacturer's ratings.

If you have a transformer on which you have no specs. except that it is designed to couple from a low impedance to push-pull grids, a grid, push-pull plates of a plate or just "high impedance", connect that winding to the crystal diode and experiment with connecting the headphones to the various taps provided on the low impedance winding. Do this experimentation using a weak signal and pick the connection that gives the greatest volume.

#5 Published: 10/22/99; Last revision: 03/30/2008

ARTICLE 06

A Crystal Radio Diode Detector Simulation using SPICE

A crystal radio set detector may be simulated in Spice by using a voltage source V1 feeding a parallel tuned circuit L1|C1 through a source resistance R1. The parallel tuned circuit may be made to have any Q by placing a parallel resistor across the tuned circuit. In the simulation circuit files enclosed, an infinite Q is assumed (no RF tuned circuit losses). The actual source loaded Q of the tuned circuit is R1/(Reactance of C1 at resonance). The voltage at the hot end of the tuned circuit is connected through a diode D1 to a parallel RC load R2|C2. The detected output voltage is developed across this load. The purpose of doing this is to enable experimentation to determine how the detection sensitivity changes if the diode type, diode source resistance, and/or load resistance are changed. This program enabled me to develop the graphs shown in Article #1 on my home page that show how detector power loss varies as a function of rectified diode current for a HP 5082-2835 diode and also, more importantly, as a function of diode saturation current Is. The input voltage is modeled as an un-modulated 1.0 MHz sine wave consisting of 4002 individual cycles, sampled at eight points per cycle. If one wants to evaluate the result of using an AM modulated wave, three simulations can be made using min., carrier, and max. Voltage levels of the desired modulated wave. Note: Graphs of diode current and voltage waveshapes, as a function of signal power, may be viewed in Article #8.

One of the simulations in the enclosed Zip archive 'Crystal Set SPICE Simulations' (click here) uses a Spice model of a Schottky diode similar to the HP 5082-2835. This is called simulation XtlSetSim1 and its files are contained in the

directory XtlSetSim1. The other simulation uses a Spice model of the 1N34A. This is called simulation XtlSetSim2 and its files are in the directory called XtlSetSim2. Each of these directories contains all the files that were generated by my SPICE simulator when I ran each simulation. The diodes used in each of these models have the value of CJO set to 0.0 pF. This does not effect the simulation and makes it easier to experiment with various values of C2 without the detuning effect of CJO. The input source and output load resistance values are equal and match the diode RF input impedance and audio output impedance values. One would expect this condition to give the lowest loss (highest Xtal Set sensitivity) at very low signal power levels. This is not so because at very low input power levels, the diode detector exhibits a square law relation, not linear relation between output and input power. See Article #15 for an explanation of how a theoretical 2 dB increase in detector output can be obtained by a deliberate RF mismatch.

In XtlSetSim1 the input sine wave voltage is set to a peak value of 0.1 volts. Since the source resistance is set to 700,000 ohms, the power incident on the detector is -57dBm. The output power delivered to R2 is -68dBm at 10.5 mV. The scale is not shown for the green output curves in the graphs below. That scale is 0.002 mV per division with the zero depressed two divisions below the zero centerline used by the other graphs. A broadcast AM voice signal, if it developed a peak instantaneous power in the detector load of -68dBm, would be just sufficient to enable me to understand about one half the words. This assumes that I am using headphones of an equivalent 700,000 Ohms AC impedance having the power sensitivity of a good real world Sound-Powered Headset. (The 700,000 Ohm impedance, of course would be obtained with the aid of an audio transformer.) The RMS audio power would