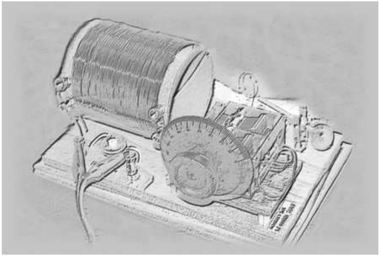
Crystal Radio Set Systems:

Design, Measurement, and Improvement

Volume II

A web book by Ben Tongue



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

NOTES:

Note: An easy way to use a DVM ohmmeter to check if a ferrite is made of MnZn of NiZn material is to place the leads of the ohmmeter on a bare part of the test ferrite and read the resistance. The resistance of NiZn will be so high that the ohmmeter will show an open circuit. If the ferrite is of the MnZn type, the ohmmeter will show a reading. The reading was about 100k ohms on the ferrite rods used here.

#29 Published: 10/07/2006; Revised: 01/07/08

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

in the Fair-Rite catalog as being above 350 degrees C. The annealing process reduces the permeability somewhat, but reduces the loss factor substantially.

The low loss-factor property of the annealed perminvar ferrite can be easily degraded by mechanical shock, magnetic shock or just physical stress (as from a tight mounting clamp). The Fair-Rite catalog sheet for type 61 ferrite cautions "Strong magnetic fields or excessive mechanical stresses may result in irreversible changes in permeability and losses". Actually, the changes are reversible if one goes through the annealing process again. The MMG catalog, issue 1A, in writing about perminvar ferrites, adds: "Mechanical stresses such as grinding and ultrasonic cleaning increase the permeability and lower the Q, especially at the higher frequencies, although the changes in Q at the lower frequencies may be very small.

I suspect that there is now much less pressure on ferrite manufacturers to deliver a low loss product than in the past. Since time is money, maybe they now skimp on the annealing process. Several years ago I took some 4" x 0.5", mix 61 rods I had purchased from CWS ByteMark and had them re-annealed at the plant of a local ferrite manufacturer. The Q of a litz-wire coil using the re-annealed core, at 2.52 MHz, was increased by 12%. This indicates that the core was not originally properly annealed, or had been subjected to some mechanical or magnetic shock after being annealing by the manufacturer. I was informed, when I asked, that coil Q at high frequencies could be expected to increase by up to100% from the pre-annealed value. I chose the best of these re-annealed rods to be my "best ferrite core" rod in this Article. One source informed me that few ferrite manufacturers perform the annealing process anymore. Toroids made of type 61 material are still made here in the USA.

ohms overwhelming the increase in DC conductor resistance from 0.16 to 2.62 ohms.

See Table 3 for measured inductance and Q values of an inductor similar to inductor BB, but wound with 125/46 litz wire. Here the Q is even greater than in Table 7 because litz construction is less sensitive to proximity and skin effect losses than is solid wire.

Thanks must go to Brian Hawes for making me aware of the FEMM program and showing me how to use it.

Part 6: Perminvar ferrite, and what the term means

Normal nickel/zinc ferrites (NiZn), the types with less permeability as well as lower loss factors than manganese/zinc (MnZn) ferrites, are often used at RF because of their low loss at the higher frequencies. They do not suffer appreciably from permanent changes in permeability or loss factor from exposure to strong magnetic fields or mechanical shock such as grinding, or dropping on the floor.

Special nickel/zinc ferrites, called perminvar ferrites can achieve a considerably lower loss factor for the same permeability than normal nickel/zinc ferrites, and at higher frequencies. This result is achieved by adding a small amount of cobalt to the ferrite power before firing, but there is a catch. In order to actually achieve the lower loss factor, the ferrite core must be annealed by raising it to a temperature above its Curie temperature (the temperature at which it losses all its permeability), and then cooling it very slowly back down through the Curie temperature, and then to lower temperatures. This process usually takes about 24 hours. The Curie temperature of ferrite type 61 (a perminvar ferrite) is specified

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maximize Q, do not cover the whole length of the core with the solenoid.

Table 7: Simulation of inductor BB in FEMM at 1 MHz, with various
conductor diameters (type 61 core material)

Wire dia. in inches	Inductance in uH	Resistive losses in ohms	Hysteresis losses in ohms	Total losses in ohms	DC resistance	Q
0.02530	258.5	11.16	1.32	12.48	0.16	130.1
0.02320	259.6	8.04	1.33	9.36	0.18	174.2
0.02127	260.5	6.26	1.33	7.59	0.22	215.7
0.01951	261.1	5.13	1.34	6.47	0.28	253.7
0.01789	261.6	4.37	1.34	5.71	0.36	288.0
0.01265	263.4	2.91	1.35	4.26	0.64	388.1
0.008995	264.0	2.48	1.36	3.84	1.25	431.9
0.006300	264.4	3.02	1.36	4.38	2.62	379.7
0.008995*	264.5	2.57	1.40	3.97	1.00	418.6

* Simulates winding the 58 turn solenoid directly on the 4" long ferrite core (solenoid ID=0.5013") instead of on a former having an ID of0.6263". Note that the the two simulations using a conductor diameter of 0.008995" show remarkably similar parameter values.

Table 7 shows the benefits of spaced winding when using solid wire. All the inductors in Table 7 use centered solenoids of 58 turns and a length of 1.624". The only variable is the diameter of the conductor, which controls the spacing of the turns (the winding pitch is held constant). The lesson here is that, when using solid copper wire, there can be a great Q benefit by space winding the solenoid and using an optimum size wire; in this case a Q of 431.9 vs 130.1 at 1 MHz, with solid wire. One can see that core losses change very little with the various conductor diameters (Hysteresis losses in ohms). Notice how, with a conductor diameter change from 0.02530 to 0.00006300", the AC copper loss decreases from 11.6 to 3.02

Part 5 - Ferrite-rod inductor simulation experiments; all using centered solenoids 1.624" long and having 58 turns

The solenoids used in the simulations in Table 6 all use a conductor having a diameter of 0.0253". The only parameter varied is the core length. The simulations in Table 7 all use a 4" long core. The only parameter varied is the diameter of the conductor.

Table 6: Simulation of solid copper wire inductor BB in FEMM at 1 MHz,
with various core lengths (type 61 core material)

Core length in inches	Inductance in uH	Resistive losses in ohms	Hysteresis losses in ohms	Total losses in ohms	DC resistance	Q
No core	17.58	1.25	-	1.25	0.16	88.4
1.68*	121.8	21.12	0.23	21.35	0.16	35.8
2.5	186.7	13.81	0.58	14.39	0.16	81.51
4.0	258.5	11.16	1.32	12.48	0.16	130.1
8.0	341.6	9.80	3.06	12.86	0.16	166.6
16.0	374.2	9.48	4.39	13.87	0.16	169.6
32.0	378.4	9.44	4.67	14.10	0.16	168.6

* Solenoid winding covers the full length of the core.

Table 6 shows that about 77% of the maximum Q is attained with a core about 2.4 times the length of the solenoid with the turns number, solenoid size, core length, etc used here. About 68% of the maximum inductance is attained. Note also that when the length of the core is shortened to approximately the length of the solenoid, Q drops precipitously. Resistive losses are mainly proximity effect losses. Hysteresis losses are magnetic losses in the ferrite core itself. Total losses are the sum of the two. There is a good lesson to be learned here: To

ARTICLE 17A

New ways to Increase Diode Detector Sensitivity to Weak Signals, and a way to determine if a diode detector is operating above or below its Linear-to-Square-Law Crossover Point

Quick Summary: The very low signal sensitivity of a crystal radio set can be improved by cooling the diode. This possibility arises when the rectified DC current is below about twice the Saturation Current of the diode. Also see Article #28 for more info on increasing weak-signal sensitivity.

Definitions of terms to be used:

- 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
- Pi Available input power
- Po Output power
- sqrt Take the square root of the expression following
- C Temperature in degrees Celsius
- Ri Detector input resistance
- Ro Detector output resistance
- R1 Source resistance
- R2 Load resistance
- I2 Rectified current
- Rxc Slope of voltage/current curve of a diode at the origin (axis-crossing resistance). Rxc=0.02568*n/Is, at 25° C.

Kt Temperature in ° Kelvin

S11 A measure of input impedance match. S11=20*log|[(|Ri-R1)/(Ri+R1)]|

SPICE A circuit simulation computer program. ICAP/4 from Intusoft was used in all simulations.

The old Article #17 has been separated into two Articles. This new Article #17A is a revision of Part 2 of the old #17. Part 1 has been broken out and renamed "Quantitative Insights into Diode Detector Operation Derived from Simulation in SPICE, and some Interesting new Equations.". It is numbered15A.

Assume that a station one can barely read has a power sufficient only to operate the detector at or below the "Linear-to-square law crossover point" (LSLCP). This is the point where the rectified diode DC current is about twice Is. Volume can be increased if the Plsc(i) point could be shifted to a lower RF power level. This will result in less insertion power loss since operation will now be closer to the linear region. The RF power required to operate a diode detector at its Plsc(i) point (at 25° C.) is shown as equation (4a) in Article #15A. It can be rewritten as:

Plsc(i)=0.0010341*Kt*Is*n Watts (1)

When referring to the schematic of a diode detector, Figure 1 will be used.

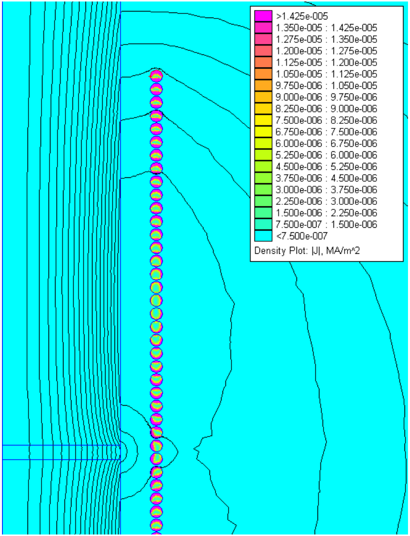


Fig. 17 GappedRodB_j: This is a zoomed in view of GappedRodB to more clearly show the current density distribution in cross-sections of the wire turns of its solenoid.

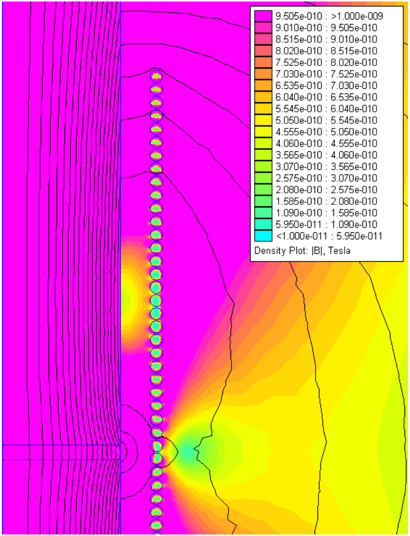
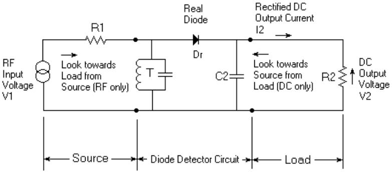


Fig. 16 GappedRodB_B: This is a zoomed in view of GappedRodB to more clearly show the magnetic flux distribution in the air and the cross-sections of the wire turns of its solenoid.



Diode Detector Schematic

Fig. 1

Diode Detector Output and Insertion Loss vs. Input Power. The LSLCP
is shown by the black arrow.

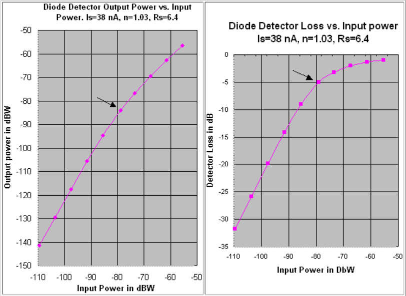


Fig. 2 - A SPICE simulation of the relation between output and input power.

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

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

It is assumed that input and output are impedance matched. One can see from equation (1) that if Is, Kt or n can be lowered, the Plsc(i) point is lowered and therefore, the volume from weak signals can be increased. The reciprocal of the product of Is, n and Kt can be seen to be a sort of "weak signal diode figure of merit" (WSDFM). It has been shown that in all semiconductor diodes, a small % drop in Kt will result in a much larger % drop in Is from its initial value. It must be remembered that the reduction of Is or Kt increases Ri and Ro. If n is reduced, Ri and Ro are reduced. Re-matching of impedances (Ri and Ro) is required to gain the benefits being sought.

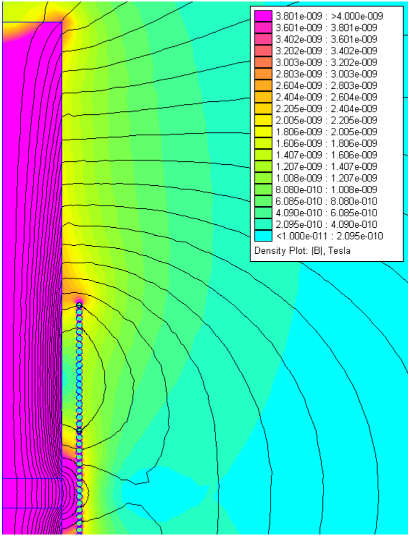

Fig. 15 GappedRodD1250: Inductance=130.011uH. Series resistive loss components: copper=8.68753 Ω, ferrite magnetic loss=0.339467 Ω, total resistive losses=9.02699 Ω. Inductive reactance=816.883 Ω. Q @ 1 MHz =90.4934.

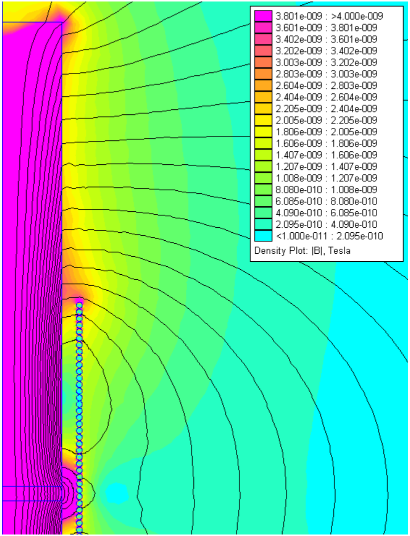

Fig. 14 GappedRodC0625: Inductance=162.729uH. Series resistive loss components: copper=6.85818 Ω, ferrite magnetic loss=0.52394 Ω, total resistive losses=7.38212 Ω. Inductive reactance=1022.456 Ω. Q @ 1 MHz =138.5044.

* Reduction of Is: The main limit to using a diode of a lower Is has to do with the resultant increase in RF input (Ri) and audio output (Ro) resistances of the detector. Practical low loss RF and AF impedance matching will be a problem. At input signal levels at or below the Plsc(i) point, those values are about: Ri = Ro = 0.00008614*n*Kt/Is ohms. The example in Figs. 2 and 3 are for a case where Ri and Ro are both about 700k ohms, using a diode with an Is of 38 nA and an n of 1.03. This is close to the limit of practicality and applicable mainly to crystal radio sets using a single tuned, high inductance, high Q loop antenna with a high quality, high transformation ratio audio transformer. A practical maximum value for R2 for most high performance crystal radio sets designed for use with an external antenna is about 330k ohms. This requires a diode with an Is of about 80 nA instead of 38 nA, for a good impedance match. The higher Is of the diode increases Plsc(i) by about 3 dB and that reduces the output of signals that are well into the square law region by about 3 dB. Signals well above the LSLCP are hardly affected at all. Note that "production process variation" of Is is usually rather great. This approach is practical and just requires selecting a diode type having the optimum Is. Simple as that, no mumbo-jumbo. See Table 1 in Article #27 for measured Is values of several diode types. Keep in mind that some diode types can be damaged by static electricity. If the diode is not destroyed, it's reverse leakage current gets elevated, ruining weak signal sensitivity. Usually, diodes that have low values of Is also have a low reverse breakdown voltage, increasing their susceptibility to static electricity damage.

* Reduction of n: The value of n does not vary as much as does Is among diodes of the same type. Schottky diodes designed for detector use usually have a low value for n. N can range between 1.0 and 2.0. Probably so called 'super

diodes' have a low n and their values of Is and n are such that a good impedance match is realized in the particular crystal radio set used. The use of a diode with a reduced n not only reduces Plsc(i), but also reduces Ri and Ro, a reverse effect than that from reducing (Is). Most diode types rated for use as detectors or mixers usually have a low n.

* Reduction of Kt: The temperature of any diode can be lowered by spraying it with a component cooler spray (221 degrees K.) every so often. A longer lasting, but lesser cooling effect can be had if the diode is placed crosswise through two diametrically opposite small holes in a small housing (such as a 1" dia. by 2.5 inch long plastic pill container) with a stack of old style copper pennies in the bottom to act as a thermal mass. This assembly is used after being cooled in a home freezer to about 0 degrees F. (255 degrees K.). It is then taken out and connected in the crystal radio set. An even lower temperature can be attained if some pieces of dry ice (195 degrees K.) are substituted for the pennies. The problem with reducing Kt is that (Is) is very temperature sensitive, so it also changes. Agilent states in App. note #1090 that the junction resistance of HSMS-2850 Schottky diode increases 100 times for a 70 degree K. reduction in temperature. That indicates a much greater % change in (Is) than in degrees Kelvin temperature. A 70 degree K. temperature drop may reduce the Is by 100 times, raising Ri and Ro by 100 times. That ruins impedance matching and increases loss greatly (the signal goes away). The answer is to experimentally try diodes that have a high Is at room temperature (298 degrees K.), that will drop to the correct value at the reduced temperature. One candidate is the Agilent HSMS-2850 (room temperature Is = 3000 nA). Another is a 2N404A Ge transistor with the base and collector leads tied together (room temperature Is = 1500 nA). Most modern diodes sold as 1N34A have (Is) values ranging from about 200 to above 600 nA. Measurements show that for

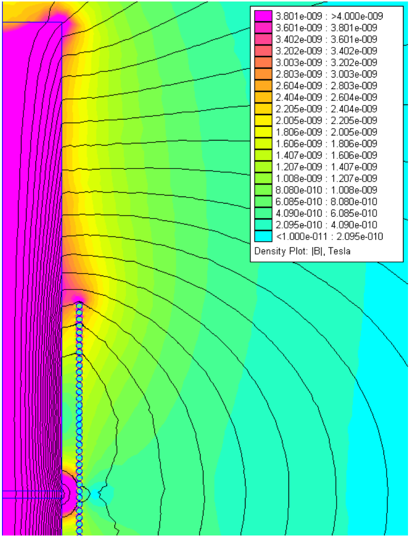

Fig. 13 GappedRodB0313: Inductance=193.324uH. Series resistive loss components: copper=6.34439 Ω, ferrite magnetic loss=0.736286 Ω, total resistive losses=7.08067 Ω. Inductive reactance=1214.69 Ω. Q @ 1 MHz =171.55. Note that this simulation has the highest Q.

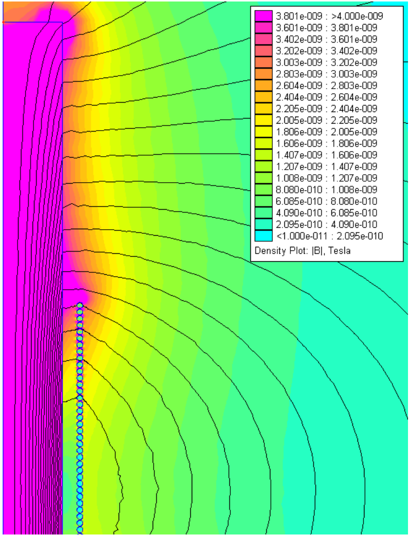

Fig. 12 GappedRodA0000: Inductance=258.539uH. Series resistive loss components: copper=11.1639 Ω, ferrite magnetic loss=1.32006 Ω, total resistive losses=12.4840 Ω. Inductive reactance=1624.45 Ω. Q @ 1 MHz =130.122.

germanium or non-zero-bias type silicon Schottkys, a 10 degree C (18 degree F.) change in temperature will result in an approximately two times change in Is. Other measurements show that with zero-bias-type Schottkys, a 14 degree C. (25 degree F.) change in temperature will result in approximately a two times change in Is. This approach is not practical since the desired results can be attained by selecting a diode type having the required Is at room temperature.

The ideality factor (n) of the diode 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 power loss change (weak signals only) 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. This illustration shows the importance of a low value for n.

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 simulation of an equivalent single diode having the same Is but an n of twice that of one original diode.

A diode detector is operating at its LSLCP (usually with about a 5 dB insertion power loss), if the average rectified DC voltage across the resistive component of its load is (n*51) mV (See Article #15 for a discussion of this). When checking this, use a large enough bypass capacitor across the DC load to maximize the voltage. If one doesn't know the n of one's diode detector, it can usually be assumed to be about 1.07. A requirement for the (n*51) mV relation to be correct is that the detector be approximately impedance matched at its input for RF and accurately matched at its output for DC and audio. Specifically, the DC load resistance must be set to 0.026*n/Is ohms (see Part 4 of Article #0 for info on n and Is). See Fig. 5 in Article #26 for a typical method of adjusting the DC resistance of the diode load and monitoring the rectified voltage. Typical values for n and Is for many diodes may be found in Articles #16 and 27. The audio load AC impedance matching requirement is not very important if one is interested only in hearing the volume delivered from ones headphones when the diode is operating at its LSLCP. The reason is that volume is a slow and gradual function of audio mismatch, for moderate mismatches. A two-to-one audio mismatch causes a loss in audio output of only 0.5 dB. A four-to-one mismatch causes a loss of 1.9 dB (hardly audible).

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coaxially, and spaced apart by 0.0000", 0.0313", 0.0625" or 0.1250". Ferrite rod diameter=0.5", Permeability of ferrite rod=125. Loss factor of ferrite at 1 MHz=30*10^-6, ID of solenoid=0.6263". Number of turns=58. Wire: solid copper, OD=0.0263". Length of solenoid=1.624". Frequency at which the simulations are made=1 MHz.

It is interesting to see in Figs. 16 and 17 that the distribution of magnetic flux in the cross-section of the turns has the same shape as that of the current density.

Please note in the text accompanying Figs. 12-17 that the copper series loss component corresponds to the sum of Ra and Ra*(FDF-1) as shown in Fig. 1.

Results of the simulations:

The first group of four simulations have central gap widths of 0.0000”, 0.0313”, 0.0625” and 0.1250”. They are named GappedRodA0000, GappedRodB0313, Gapped RodC0625 and GappedRodD0125. They are intended to show magnetic flux density distribution in the ferrite and the air. The simulations are made at a frequency of 1 MHz with an AC current of 1 uA RMS in the solenoid. The actual magnetic flux density values can be estimated by comparing the color display to the color chart to the right of the images.

A second group of two simulations shows magnified views of two parameters of the GappedRodB0313 simulation. They are called GappedRodB0313_B (for showing flux density B in Teslas in the cross-sections of the individual wire turns) and GappedRodB0313_j (for showing the current density j in MA/m^2 in the cross-sections of the individual wire turns).

If one looks closely at the GappedRodB_B image, one can see how flux density is distributed in the wire cross-sections as a function of distance along the rod. As the textbooks say, very little flux exists in the interior of the wire. Where the external flux density is great, as it is at the ends of the rod and near the gap, the flux that penetrates the copper is confined near the outer periphery of the wire.

The distribution of current density in the turns as a function of position along the length of the rod is shown in image GappedRodB0313_j. This illustrates skin effect. Note the current density is not uniform in the wires because of proximity effect and the fact that the length of the solenoid is not very long, compared to the length of the rod.

Some specifications common to the inductors in Figs. 12 through 17: Ferrite rod length: two 2” long rods oriented

ARTICLE 18

Get 3 dB more Output for Greater Volume on Strong Stations plus...

Quick summary: Over the years many experimenters have realized that one could get "free" power from a crystal radio set and operate an amplifier with it. This has been successfully done by coupling an additional tuned circuit and detector to the antenna and tuning it to a strong station. The rectified DC from the station was then used to power an amplifier for boosting the audio output of the crystal radio set without appreciably affecting its normal operation, when tuned to a different station far enough removed in frequency. This Article describes, possibly for the first time, a method of using the carrier power of the station being received to power an amplifier.

One third the total power in a 100% modulated AM signal is in the sidebands that carry the audio modulation. An ideal, 100% efficient Crystal radio set will convert all of the received sideband power to audio output power. Call it audio power output #1. What about the other two thirds of the power? That is the power in the AM carrier that carries no audio information but has twice the power of the sidebands (at 100% modulation). This Article shows the circuit of a device that can be used to extract that carrier power and use it to operate a micro-power op-amp. The op-amp uses the detected audio voltage from the diode detector for its input and provides an additional source of audio power. Call it audio power output #2. These two audio power sources, #1 and #2 can be added together to create a final output at least 3 dB more than the normally available audio power output #1.

1. Background

Within the last year or so, Burr-Brown (now owned by Texas Instruments) came out with a micro-power op-amp (OPA349) specified to work with as little as a 1.8 volt DC power source. It draws a minuscule 1 uA quiescent supply current. This op-amp opens the possibility of building a device I call a "Free 3 dB Detector Load" (F3dBDL). I have found that the F3dBDL will actually operate with an input signal low enough to generate a rectified voltage as low as 1.2 volts DC. Maybe all the OPA349s will work in this circuit at 1.2 volts. My F3dBDL requires a minimum input carrier power of -53 dBW and a rectified DC voltage of at least 1.2 volts.

2. A conventional diode detector with standard output loads (DC and audio)

Any crystal radio set that uses an audio output transformer can be represented by the simple circuit shown in Fig. 1. V1, R1 represent the antenna-ground power source, impedance transformed to the tank circuit. The detected carrier power is dissipated in the resistive load R2. The detected side-band power is delivered to the audio load R3.

out and has little effect on Q, showing a Q ratio between the two of about 0.97. Compare these figures with those in Table 3 for a similar conventionally wound solenoid using close-spaced 125/46 litz wire. Proximity loss is greatly reduced in close-wound litz wire, compared to close-wound solid copper wire. The Q ratio here is about 0.75. Loss in the ferrite core swamps out the much lower proximity loss in the litz wire, and a much higher Q results.

Part 4 - The effect on impedance parameters of an air gap in the center of a ferrite rod inductor and on magnetic flux/current density in the wire cross-sections.

The images and text are based on FEMM simulations of inductor BB, shown in Figs. 6 and 7, the specs of which are shown below Table 4, but with the following difference: Instead of using one 4” long ferrite rod, two 2” co-axially oriented rods are used in each simulation with different air gaps between them (0.0000”, 0.0313”, 0.0625” and 0.1250”). The solenoid is centered on the gap. When the gap is 0.000” the result should be the same as if one solid 4” rod were used. The inductance and Q values for inductor BB are slightly different than the values in the new simulations, as stated below for a 0.0000” gap between two 2” rods. This is because the “meshing” parameter in the FEMM simulation was changed to reduce the time taken for the simulations.

The top half of a full image of a core/solenoid combination is a mirror image of the bottom half. In figures 12-17 advantage is taken of this characteristic by zooming in and not showing the entire rod so as to be able to get a larger image, thus supplying more detail.

wire, OD=0.0253", solenoid length=1.624", ID of solenoid=0.5013", Number of turns=58, Inductance=261.66 uH, Q=118.4. Solenoid construction is similar to inductor A in Tables 1 and 2.

*

Inductor BB: ferrite core length=4", core diameter=1/2", core type=61, wire type=22 ga. solid copper wire, OD=0.0253", solenoid length=1.624", ID of solenoid=0.6263", Number of turns=58, Inductance=259.11 uH, Q=130.7. Solenoid construction is similar to inductor B in Tables 1 and 2.

*

Inductor DD: Same as inductor AA except that the wire diameter is reduced to 0.01765". This creates a spaced winding. Inductance=265.37 uH, Q=267.6.

*

Inductor EE: core length=1.680", core diameter=1/2", core type=61, wire type=22 ga. solid copper, wire, OD=0.0253", solenoid length=1.624", ID of solenoid=0.6263", Number of turns=58, Inductance=121.80 uH, Q=36.2.

Table 5: Measurements at 1 MHz of a physical inductor having the same parameters as simulated inductor BB. Inductance=~236 uH

--	**"Best core"**	**"Worst core"**
Frequency in Hz	**Q**	**Q**
540	130	127
943	141	138
1710	150	145

Note that the Q difference between the "Best core" and the 'Worst core" is very small. This is because the main loss in this inductor is the high proximity loss in the solid close-spaced copper winding. The much lower ferrite core loss is swamped

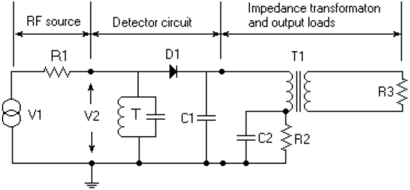


Fig. 1 - Equivalent Circuit of a Conventional Crystal Set

Schematic of crystal set with conventional load

3. Conventional crystal radio set detector with the F3dBDL

The F3dBDL is intended to be used with signals strong enough to cause the detector to operate in its peak-detection mode. In this case, the DC load R2, seen by the diode D1 should equal to two times the RF source resistance R1. D1 should also see an AC load resistance of two times R1, at the primary of transformer T1. (See Article #0 , Section 4, for more info on this.) The power dissipated in the DC load R2 in the circuit in Fig. 1 will be used, in Fig. 2, to power the op-amp U1. In Fig, 1 the audio output power is delivered to the output load R3. In Fig. 2 audio output power is delivered to two loads of value R3 and k*(R3). With proper selection of the relative impedance transformation ratios of T1 and T2, the value of k may be made equal to about 1. In addition, the output currents of T1 and T2 become about equal. In this case, no current will flow in connection X, and it can be eliminated. This gives us one 600 ohm instead of two 300 ohm outputs. The resultant load resistance of twice R3 will absorb twice the audio power than did R3 in Fig. 1, although at twice the impedance (600 ohms). The resistive network R4, R5, R6 and R7 biases + input terminal of U1 at 1/2 the DC supply voltage appearing across C2 and attenuates the detected audio voltage appearing across C1 so that it will not overload U1. The value of capacitor C2 is made quite large to enable it to hold steady the voltage it supplies U1, between bursts of speech.

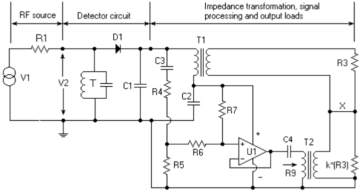


Fig. 2 - Circuit of Fig. 1 with its Output Loads Replaced by the Free 3 dB Signal Processing Circuit and its Output Loads (F3dBDL)

Schematic of crystal set with active load

Parts List

Table 4 - Simulation of inductors using solid copper wire of OD=0.0253" in Figs. 4, 5, 6, 7, 10 and 11. Wire OD=0.01765" in Figs. 8 and 9. No litz wire is used. All inductors have 58 turns.

Fig. 4 Simulation of inductor AA

Fig. 5 Close-up view of flux density near upper turns of inductor AA

Fig. 6 Simulation of inductor BB

Fig. 7 Close-up view of flux density near the upper turns of inductor BB

Fig. 8 Simulation of inductor DD, same as AA except for using wire of a smaller OD

Fig. 9 Close-up view of flux density near the upper turns of inductor DD

Fig. 10 Simulation of inductor EE, short ferrite core

Fig. 11 Close-up of flux density near the upper turns of inductor EE

Parameters of simulated inductors AA through DD, inductance and Q at 1 MHz:

*

Inductor AA: ferrite core length=4", ferrite core diameter'1/2", core type=61, wire type=22 ga. solid copper

Table 4 - Simulation of inductors using <u>solid copper wire</u> of OD=0.0253" in Figs. 4, 5, 6, 7, 10 and 11. Wire OD=0.01765" in Figs. 8 and 9. No litz wire is used. All inductors have 58 turns.

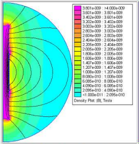


Fig. 4 Simulation of inductor AA

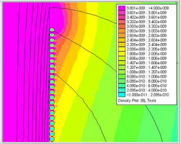


Fig. 5 Close-up view of flux density near upper turns of inductor AA

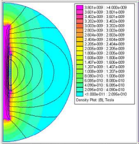


Fig. 6 Simulation of inductor BB

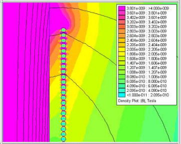


Fig. 7 Close-up view of flux density near the upper turns of inductor BB

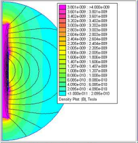


Fig. 8 Simulation of inductor DD, same as AA except for using wire of a smaller OD

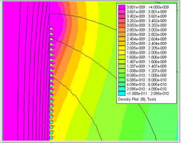


Fig. 9 Close-up view of flux density near the upper turns of inductor DD

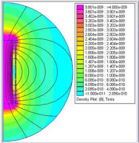


Fig. 10 Simulation of inductor EE, short ferrite core

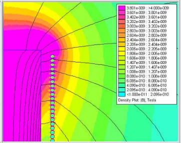


Fig. 11 Close-up of flux density near the upper turns of inductor EE

C1	47 pF - RF bypass. Physically, it will probably be supplied by the winding capacitance of T1.
C2	10 uF low leakage electrolytic - voltage holding storage capacitor for the supply voltage of U1
C3	1nF - audio coupling capacitor
C4	10nF - DC blocking capacitor
D1	1N34A or several Agilent 5082-2835 or HSMS2820 in parallel
k	A constant, which when multiplied by R3 gives the value of the load on T2
R1	Transformed source resistance of antenna across tank T - assumed to be 150k ohms
R2	DC diode load resistance in Fig. 1 - Assumed to be 300k ohms
R3	Audio load resistance in Fig. 1 and 2 - Assumed to be 300 ohms
R4	2.2 Meg resistor
R5	5.1 Meg resistor
R6	5.1 Meg resistor
R7	10 Meg resistor
R9	AC resistance seen looking into the primary of T2 when connection X is present
T1	Transformer with 1000:1 transformation ratio - such as Stanley 100k to 100 ohm unit, available from Fair Radio Sales as part # T3/AM20, or UTC C-2080
T2	Transformer with 100:1 transformation ratio - such as Calrad 45-700, available from Ocean State Electronics.
U1	Burr-Brown Opamp #PA349 - available from a Texas Instruments distributor. A convenient way to connect to the tiny leads of IC1 is to first solder it to a surfboard such as one manufactured by Capital Advanced Technologies (http://www.capitaladvanced.com). Their models 9081 or 9082 are suitable and are available from various distributors such as Alltronics, Digi-Key, etc.
V1	Internal voltage of RF power source (antenna induced voltage after impedance transformation to the tank circuit "T")

4. Comments

If all the power in the carrier could be changed to audio power and added to the main detector audio output, the total audio power would be tripled, a 4.8 dB increase. It would be nice if the op-amp had 100% efficiency in converting its input DC power to output audio power, but it doesn't. An ideal class B amplifier has a theoretical efficiency of 78.5%. Therefore, we lose at least 21.5% (1.05 dB) right off the bat. Other losses in the op-amp, the 1 uA quiescent current of the U1 and the bias network R5, R6 and R7 use up some more of the 4.8 dB. The transformer T1 has losses and uses up some more of the 4.3 dB. We are left with an output power from the U1, T2 combination about equal to that of a conventional crystal radio set. The two added together gives the 3 dB increase.

There are some limitations in using the F3dBDL. The IC is specified to operate over a supply voltage range of 1.8 to 5.5 volts. In this circuit it seems to work well over a supply voltage range of 1.2 to greater than 5.5 volts. This corresponds to an input carrier power range of -53 to >-40 dBW. I have found, that for me, the volume to be too great for headphone use but barely adequate for high efficiency horn speaker use. If more than -40 dBW of AM signal carrier power is available, the F3dBDL can be made to handle it (and give a greater sound volume) if the F3dBDL is operated at a lower output impedance level. In this case, transformers T1 and T2 might have to be changed to ones with a lower transformation ratio.

The impedance at the + signal input terminal of U1 is very high. Use care to minimize stray capacitance to ground at this point. Too much will roll off the highs. The high audio frequency output capability of U1 falls as signal strength and, as a result, supply voltage increases. This can cause audio distortion.

The F3dBDL can also be used to increase the volume on weak stations. This is done by connecting a ceramic electric double layer high capacitance capacitor across C2, charging it up overnight on a strong station and then switching it to power the opamp for weak station listening later on. A 0.047 Farad capacitor will hold its charge for many hours in this application. One manufacturer of this type capacitor is Panasonic, and one of their distributors is Digi-Key Corp.

If the load on the F3dBDL is a SP headphone set with the elements wired in series, bass response can be improved with a

simulation. It's made up mostly of air. The magnitude of the flux density can be seen from the colors on the display (see the chart). The range of flux density values for the display was purposely limited to about 20 to help supply flux density detail around the outer turns of the solenoid. That is why most all the core is colored purple (the flux density is above 4.000e-9 Tesla). Fig. 5 is a close-up simulation of the area near the upper turns of the solenoid. If one's browser has a zoom control, one can easily see how the flux density close to the surface of the wires of the end turns of the solenoid (even numbered Figs.) is greater than it is in the more central turns. High flux density in the copper equals high power loss (Q reduction).

Comment: Look at figs. 6 and 11 in Table 4. Inductors BB and EE are identical except for the length of the ferrite rod. It appears that about 10% of the end turns of solenoid BB are exposed to a flux density above 2.8e-9 Tesla (3 dB below the maximum plotted value of 4e-9 T). The corresponding percentage in solenoid EE about 50%. This shows that a high flux density around a greater percentage of turns results in lower Q. A parameter listing of the inductors is below Table 4. Note the Q values for inductors BB and EE.

David Meeker's "Finite Element Method Magnetics" program FEMM was used to generate Figs. 4-17. First a word about the displays: FEMM, as used here provides a 2-dimensional display of flux density (the colors) and flux lines (the black lines). Only half of the object being simulated is analyzed and displayed since only axisymmetric objects can be analyzed with the program. This saves simulation time, which can become very great. FEMM, at this time, cannot simulate using litz wire. That is why the following simulations and measurements use 22 ga. solid copper wire instead of the 125/46 litz used in Part 2.

Understanding the images: Visualize the axis of the ferrite rod as coincident with the y-axis of a conventional 3-D x, y and z coordinate graphing system with the center of the rod at the origin. The FEMM program discards everything to the left of the y, z plane that intersects the origin and displays a view of the other half. The images show field densities that exist in an x, y plane that intersects the origin. The vertical object at the left in each image is the ferrite rod. Its horizontal width is ½ that of the diameter of the actual round ferrite rod since the parts of the inductor to the left of a vertical y, z plane intersecting the origin have been discarded (see above). The vertical line of little circles to its right show the cross-sections of the turns of the solenoid wires. Fig. 4 is a plot of magnetic flux density and flux lines on an imaginary plane that cuts longitudinally through the center of ferrite rod inductor AA, shown mostly in purple. The outline of half the 4"x1/2" rod is shown at the left of the plot. If one measures, on the computer screen, the height and width of the rectangle, one can see that their ratio is 16. This is equal to the ratio of the 4" length of the rod to 1/2 of its 1/2" diameter. The large half-circle defines the area around the inductor that will be included in the

small subjective increase in volume. Consider the two headphone elements as the two impedance equal loads R3 and k*(R3), in Fig. 2. Restore the connection X. The element k*(R3) will have a much better bass response than the other one because it is driven by the low output resistance of the opamp. See "It is interesting to note" at the end of Section 1 in Article #2 for more info on this.

Last, but not least, one should not expect too much from the F3dBDL. After all, a 3 dB or so increase in volume will not be perceived as a lot. The challenge of this project was to devise a way to use all of the power in an AM modulated signal, I believe that has been accomplished.

#18 Published: 07/09/01; Revised: 04/06/2007

ARTICLE 19

An Explanation of how the "Mystery Crystal Radio" Works

Quick summary: Plans for a crystal radio called the "Mystery Crystal Set" were published in the newspaper "The Sunday Mail" of Brisbane, Australian in 1932. The "Mystery" in the name comes from the fact that, in the schematic, there seems to be no ground return to which the antenna currents can flow. The design was used by entrant Ray Creighton in the "Crystal Set Competition" held on March 19 2000 by the Southeast Queensland Group of the Historical Radio Society of Australia in Malaney, Australia . His entry won first prize in one category and third prize in another. see it here The design has recently become popular in the US as shown by the many messages posted on the Yahoo! Groups site "thecrystalsetradioclub". On 6/6/2000, in messages 2172 and 2173, I posted the following explanation (edited here) of how I believe the Mystery Set works:

length? If one wishes to use solid wire, it should probably be wound directly on the ferrite, not on a former. The overall Q will still be much less than when using litz, and the loss from the high (tan δ) dielectric of the ferrite will be pretty well swamped out because of the now higher losses from the skin and proximity effect losses. The Q values, using a close-wound solenoid of 22 ga. solid copper wire on a polyethylene former, as in 'Coil and Former' B in Table 2 are: 520 kHz: 130, 943 kHz: 141 and 1710 kHz: 150 when using the "best core". The Q drops only 3, 3, and 5 points respectively if the "worst core" is used.

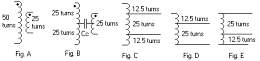
Measurements to determine the (tan δ) of the dielectric of a 'medium core': Two adhesive copper foil coupons, 0.5"x1.75" were affixed to the 4", 0.5" diameter rod made of ferrite material 61 (3M sells rolls of thin copper foil with an adhesive on one side). The long dimension of each coupon was parallel to the axis of the rod with the two coupons set opposite to each other, 180 degrees apart. They formed a two plate capacitor having curved plates with the dielectric of the rod between them. The capacitance of this capacitor came out to be 6.5 pF. Measurements, using a Q meter and a high Q inductor were made that enabled calculation of the Q of this 6.5 pF capacitor. Q was 25 at 520 kHz, 35 at 943 kHz and 55 at 1710 kHz. Even though the distributed capacity of a ferrite rod inductor is only made up partially of this poor dielectric, it is, I believe, a previously unrecognized cause of the usual Q drop at the high end of the band. It is also, I believe, the cause of Q reduction in ferrite toroids when no gap is provided between the start and finish of the winding.

Part 3 - Flux density and flux line simulations, inductance and Q of several ferrite-cored inductors along with some measurements

The Q values given above were measured on an HP 4342A Q meter and corrected for the distributed capacity of the inductor (Co). The ferrite cores were purchased from CWS ByteMark in the third quarter of 2002. They may have changed vendors since then because some rods I purchased in the 3td quarter of 2004 resulted in lower Q coils than the values reported here. These rods also had two small, 180 degree apart, longitudinal flats along their entire length. CWS gracefully accepted a return of those rods and quickly refunded my money. The 'best' and 'worst' cores used in these measurements were from a group purchased from CWS ByteMark in the 3rd quarter of 2002.

Note the better high-band Q values recorded for the contra wound inductor. This is because the low Q distributed capacity from the dielectric of the ferrite (Co and Ro) is connected across an inductor having 1/4 the inductance (and reactance) value of the conventional wound solenoid. An observation: If the hot/cold connections to the contra wound coil in Fig. 3 are reversed, Q at 1710 kHz drops. This is because more loss from the low Q dielectric of the ferrite is coupled in to the stray capacitance.

Solid wire instead of litz?: Keep in mind that the work described here used close-wound 125/46 litz wire. If one duplicates 'Coil and Former B' in Table 2, except using 22 ga. solid copper wire (having the same diameter) as 125/46 litz, the Q values drop to about 1/6 of the values achieved with the litz wire. The cause is the large proximity effect resistive losses, as well as skin effect, in the solid wire. The proximity effect, but not the skin effect loss may be much reduced if the wires are space-wound. New trade-offs now must be considered: Same wire diameter, and therefore a longer solenoid, or a smaller wire diameter and the same overall

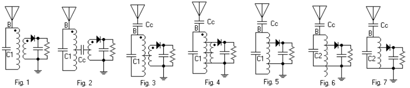


The "B" antenna connection is the one that gives the most sensitivity and the least selectivity.

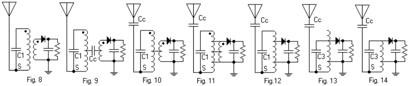
The "S" antenna connection is the one that gives lesser sensitivity and more selectivity.

These notations are the reverse from those shown in the original Mystery Set article.

Various Stages in the Simplification and Reduction of the Inductor



Simplification and Reduction of the Circuit when using the "Broad" Antenna Connection



Simplification and Reduction of the Circuit when using the "Selective" Antenna Connection

Equivalent Schematic analysis of the Mystery Crystal Set

Two assumptions made in the analysis: They are that the distributed capacity between the two coil windings may be represented by one lumped capacitor, Cc, connected between the center of one winding to the center of the other. See Fig. B. The other is that the magnetic coupling between the primary and secondary windings is very high. This assumption is close to reality for the bifilar wound portion of the transformer, provided the capacity coupling is not too high. The magnetic coupling between the bifilar-ed parts and the end windings is not as close as that between the bifilar-ed parts. This does not affect the validity of the analysis. Keep in mind that in transformers with unity coupling, the ratio of the voltage on any winding to any other is directly proportional to the number of turns on each winding. This also applies to a portion of one winding. (Just use the number of turns in that portion.) Figures A through E show the inductive circuit through various changes as the following reduction and simplification proceeds.

Simplification and reduction of the circuit of the Mystery crystal set using the "Broad" non-earthy antenna connection:

The physical circuit of the Mystery set is shown in Fig. 1 with the antenna connected to the non-earthy side of the primary. The black dots on the windings show the start of each winding, assuming that they are both wound in the same direction.

Figure 2 shows a coupling capacitor Cc, between the two windings. It represents the parallel combination of two distributed capacitances: One is formed of the dielectric of the wire insulation between the bifilar-ed primary and secondary coil turns. The other is also between the primary and secondary coil turns, but in this case, there are three dielectrics in series. They are: (1) The dielectric of the insulation on, say, the primary winding that is in contact with the coil form. (2) The dielectric of the coil form between the primary and secondary windings. (3) The dielectric of the insulation on the secondary winding that is in contact with the coil form. Cc is in a series circuit with the antenna and ground.

Fig. 3 shows Cc shifted up to the antenna and out of the way. No change in performance will result.

The top and bottom leads of the secondary are connected (each 12.5 turns from the center), to the corresponding points on the primary (12.5 turns up and down from the center). This is shown in Fig. 4. Since the points that are connected together have the same AC voltage on them, no current will flow through their connection and the circuit operation will be undisturbed.

Ferrite rod length=4", Diameter=0.5", Material=type 61, µi=125, Ferrite loss factor (FEF)=30*10^-6, Former=polyethylene (not vinyl) tubing, ID=0.5", OD=0.625", length=5", Wire=125/46 ga. litz, Construction=close-wound contra wound solenoid of 58 turns and length of 1.625" (not wound as tightly as the conventional solenoid above).

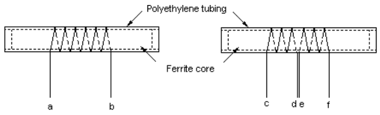


Fig.2 - Conventional solenoid, 6 spaced turns
Fig.3 - Contra-wound solenoid, 6 spaced turns

Winding format for conventional and contra wound solenoid

The winding format for solenoids #1 and #2, below, are shown in Figs. 2 and 3. For clarity, the windings are shown as space wound, but the actual solenoids #1 and 2 close wound. Connections for the contra wound inductor shown in Fig. 3: For the series connection, join leads c and e. Lead d is hot and lead f is cold. For the parallel connection, join leads c and f. Join leads d and e. d/e is the hot and c/f is the cold connection.

Table 3 - Conventional vs contra wound Ferrite-Rod Cored Solenoids

	Table 3 - Conventional vs contra wound Ferrite-Rod Cored Solenoids											
	#1 Conventional solenoid						**#2 contra wound solenoid**					
	'Best core'			**'Worst core'**			**'Best core'**			**'Worst core'**		
Freq. in kHz	Q	Ind. in uH	Co in pF	Q	Ind. in uH	Co in pF	Q	Ind. in uH	Co in pF	Q	Ind. in uH	Co in pF
520	960	237	2.8	740	240	2.8	895	231	4.0	700	234	4.0
943	~1030	237	2.8	775	240	2.8	990	231	4.0	765	234	4.0
943	-	-	-	-	-	-	~1030	57.8	4.7	780	58.5	4.7
1710	855	237	2.8	655	240	2.8	945	57.8	4.7	725	58.5	4.7

that used in 'Coil and Former' B, above. The winding had 39 turns, close wound, of 270/46 litz. The "best ferrite core" was selected from a small batch of cores that were re-annealed by a local ferrite manufacturer. See the third-from-last paragraph.

Some observations:

1. Inductance does not change much between a solenoid diameter of 0.5" and 1.5".
2. At low and medium frequencies, Q is the highest when the wire is wound directly on the ferrite. It drops substantially at the high frequency end.
3. Q at the high frequency end increases as the wire is separated farther from the core, except for coil E.
4. Q at the low frequency end decreases as the coil wire is separated further from the core.

Comparison between a conventional and contra wound ferrite-rod cored solenoid using a "best" and a "worst" rod.

See Article # 0, Part 12 for a mini-Article about the benefits of the contra-coil construction.

Ferrite rod length=4", Diameter=0.5", Material=type 61, µi=125, Ferrite loss factor (CLF)=30*10^-6, Former=polyethylene (not vinyl) tubing, ID=0.5", OD=0.625", length=5", Wire=125 strand/46 ga. litz, Construction=close-wound conventional solenoid of 58 turns having a length of about 1.625"

Figure 5 shows the resulting equivalent circuit from the connections made in #4. Since all portions of the winding are assumed to be unity-coupled to each other, performance will not change if the tuning capacitor C1 is connected as shown in Fig. 6, as long as its value is changed appropriately. C1 is connected across 50 turns of the inductor. C2 is connected across 37.5 turns. The inductances of a unity coupled 1:1 transformer are directly proportional to the square of the number of turns. The number of turns across which C2 is connected is 3/4 of the number of turns turns across which C1 is connected, therefore, the inductance across which C2 is connected will be 9/16 the inductance across which C1 is connected. C2 must be increased from the the value of C1 to 16/9 of C1 for the circuit to work the same as before the transformation. The bottom portion of the coil in Fig. 6 can be eliminated since nothing is connected to it.

The final result is the equivalent circuit shown in Fig. 7. Here we see a conventional crystal set circuit with the antenna-ground components connected directly across the full tank, with isolation from full antenna resistive loading supplied by the capacitor Cc. The detector load is tapped in at 2/3 of the tank voltage to reduce its resistive loading effect on the tuned circuit. That's it for the non-earthy "Broad" antenna connection.

Simplification and reduction of the circuit of the Mystery crystal set using the "Selective" earthy connection.

Figures 8 through 14 show the simplification and reduction of this circuit. It proceeds in an manner similar to the one for the "Broad" connection. Now look at Fig. 14. The value of Cc is unchanged from that in Fig. 7. C3 will have to be somewhat larger than C2 was for the circuit to work the same. The antenna-ground components and Cc are now connected across only 1/3 of the tank instead of the full tank. The detector load is still tapped in at 2/3 of the tank voltage. That's it for the earthy "Selective" antenna connection.

What might the value of the magnetic coupling coefficient between the bifilar-ed portion of the windings be?

To think about this, consider: Mentally unwind the bifilar portion of the coil from the coil form, but imagine the two wires are still in the same relative positions to each other. Stretch them out. The ends of one wire are the terminals of one winding of a transformer and the ends of the other winding, the terminals of the other. Now you have two parallel wires closely spaced and several tens of feet long. The spacing (from the wire insulation) between them is maybe 0.005". It should seem obvious that the magnetic coupling between them could not get much greater (without ferrite cores), no matter what one does with the wires. It can, however become greater when the bifilar wire is wound on a form. The reason is that places a primary wire on each side of every secondary wire and vice-versa, providing more magnetic coupling between the windings than when the wires are stretched out.

Here is an approach for determining the coupling coefficient of a bifilar winding: Construct a bifilar wound coil that has about the same inductance as the bifilar-ed wires in a standard Mystery" set. This inductance calculates out to be 57 uH. No wire of the gage originally used was available, so the largest bonded bifilar wire I had available was used. It was made by MWS Wire and consisted of two #30 ga. film insulated wires bonded together. Its cross section measures 0.012x0.024".

Coil and Former >	A	B	C	D	E
Coil former dia. and len. in inches	0.5	0.625x4.5*	~0.75x4.5**	1.04x2.25	1.50x3.0
Coil former Material	No former- wire wound directly on ferrite rod	Polyethylene sleeve, 1/16" wall thickness	Split polyethylene tubing placed over former 'B'	Orange colored polypropylene pill bottle	White Polypropylene drain pipe from Genova Products
Ind. of coil in air, in uH	?	19.3	25.9	48	90
Q of coil, air core, 2520 kHz	?	265	320	430	520
Inductance of coil in uH with 'best' ferrite rod	237	248	238	240	241

Table 2 - Q of a ferrite-cored conventionally-wound coil of fixed length and number of turns as a function of its diameter (uses 'best ferrite core')

Coil and Former >	A	B	C	D	E
V Freq. in kHz V	Q	Q	Q	Q	Q
520	1060	960	945	820	670
943	1035	1030	1045	995	890
1710	780	855	878	885	845

* Piece of polyethylene tubing having an OD of 0.625" and an ID of 0.50"
** This coil former has a cross section somewhat less than from a full 0.75" piece of tubing. It is constructed by first sliding the 1/2" dia. 4" long ferrite rod into a 5" long piece of 0.625" OD polyethylene tubing. A full longitudinal cut is then made in a second piece of similar tubing, so it can be fitted over the first one. A gap of about 3/8" is left in the second, slit piece of tubing, and that is what causes the cross section to be less than that of a true 3/4" tube.

Note 1: Q values are corrected for distributed capacity.
Note 2: 'best ferrite core', 'medium ferrite core' and 'worst ferrite core' refer to Q measurements of a large quantity of 4" long, 1/2" diameter ferrite 61 cores purchased from CWS Bytemark over a period of years. The Q measurements were made at 1710 kHz with a test coil wound on a former similar to

LFEF, which also provides an intermediate value for real-world inductance. See Table 6, next to last entry.

In my experience with 1/2" diameter ferrite 61 rods aiming for 250 uH and using Litz wire, most of the time LFEF turns out to be greater than the value for maximum Q. An indication of this condition can be obtained by placing two extra cores, each co-axially aligned with the solenoid's core, one at each end of said core, to increase the LFEF. The Q is usually reduced even though the inductance is increased, showing that LFEF is too high for maximum Q. Proof of this can be attained by discarding the two extra cores and reducing the number of turns on the rod. Of course, inductance goes down, but Q will increase. To get the inductance back up and retain the higher Q, a solenoid and ferrite rod of larger diameter are required.

** Note the "no core" entry in Table 6 for inductor BB. The solenoid (with no core) has a Q of 88 (and an inductance of 17.6 uH).

Part 2: Measurements

Comparison of several conventionally wound Ferrite-cored solenoids having the same winding length and number of turns, but different diameters

Ferrite rod length=4", diameter=0.5", material=type 61, µi=125, ferrite loss factor (CLF)=30*10^-6, the "best ferrite core" was used, former=low loss thin wall tubing of various lengths, wire=125 /46 ga. litz, construction=conventional close wound solenoid of 58 turns having a length of about 1.625".
Table 1 - Coil and Former data (uses 'best ferrite core')

Twenty turns were wound on an available 3 1/2" styrene coil form since a 3" diameter coil form, as used in the original Mystery set was not available. The winding length came out to be a very small 0.475" because of the small wire size. This is much less than that in the original Mystery set but, tough, that wire is all that was available. The leads from the coil were still bifilar-ed, 10" long ends.

Several resonance measurements were then taken using a Q meter. The first was with one winding connected to the inductance terminals of the Q meter, the other winding being open circuited (Loc), at several frequencies from 0.515 to 2.36 MHz. The indicated capacitance readings on the Q meter were noted. The other was with the same winding still connected to the inductance terminals of the Q meter but with the other winding shorted (Lsc), at frequencies from 3.0 to 11.0 MHz. Again, the indicated Q meter capacitance readings were noted. At frequency extremes these readings will be distorted by the presence of distributed capacitance between the two windings, 1020 pF in this case. The conventional Mystery set would have considerably less capacitance between the windings because of the much thicker insulation on the wires. Note: The capacitance between the windings cannot be determined at RF by the use of a Q meter. It can be measured by the use of an RLC bridge operating at 1 kHz or a DVM having a capacitance measuring function (if it operates at about 1 kHz).

Over the frequency range of 0.515 to 1.71 MHz, Loc was calculated to be: 66.5 +/- 2.5 uH. Over the frequency range of 3 to 7 MHz, Lsc was calculated at: 2.01 +/- 0.06 uH. A derivation results in the following relation for the coupling coefficient between two identical magnetically coupled inductors: k=sqrt(1-Lsc/Loc). The calculated coupling coefficient between the two bifilar-ed windings is 0.984, which I consider very close to unity.

The bifilar wire was re-wound on the same form, but spaced to cover a 1" length. The coupling coefficient came out at 0.966 and the distributed capacitance: 895 pF. Another coil was then wound from the same piece of wire on a 1.5" diameter polypropylene form. The winding was slightly space wound and had a length of 1.5 ". Coupling coefficient: 0.983 and distributed capacity coupling: 945 pF.

Of course, manufactured, bonded, bifilar wire is not recommended for use in a Mystery set. Usually two independent, insulated wires are wound close spaced. This practical case results in substantially less distributed capacitance than when using bonded wires.

Conclusion:

The beauty if the Mystery set is that it provides an antenna decoupling capacitor (Cc) (made from the distributed capacity between the bifilar-ed windings), along with the effect of two different points for its connection to the tank; all without any specific physical capacitor or taps on the inductor. Further, the diode is effectively tapped 1/3 down on the tank for improved selectivity. The only downside to this arrangement is some loss caused by the probable relatively low Q of Cc.

When using the "Broad" antenna connection, the antenna-ground components are connected through Cc across the full tank. This arrangement puts a relatively large amount of antenna resistive loading on the tank. The loading results in as reduced selectivity, but stronger signal strength than one gets in the "Selective position. See Fig. 7.

the coil. Generally the core providing the least inductance will provide the highest Q.

Comments: Consider the schematic in Fig. 1. La, Ra and Ra*(FDFa-1) define the inductance and Q of the air-cored solenoid (Before a ferrite core is inserted in an air-cored solenoid, FDF=1).

Lp and Rp define the inductance and Q of the added inductance produced when a ferrite core is inserted into the solenoid (now FDF becomes FDFf greater than before because of greater flux density in the conductors). The value of Lp depends upon ui of the ferrite material, La and LFEF. Some methods of changing LFEF are: 1) Increase the amount the bare rod core extending beyond the solenoid. This will increase the value of LFEF and consequently the value of Lp. 2) Use a smaller diameter ferrite core than the Id of the solenoid. This will reduce the value of LFEF and consequently Lp.

The L and Q values of the air-cored solenoid are usually quite low**. The inductance of Lp is usually high and equal to LFEF*ui*La. The parallel resistance Rp equals (reactance of La)*LFEF/CLF. If LFEF equals 1 (This can be approached when using a toroid having a high permeability, ui), the Q of a real-world ferrite-cored toroid inductor is about: Q=1/(ui*CLF). The Q of a ferrite-cored toroid inductor using ferrite 61 as the core can have a Q of about 330 at 1 MHz, as shown in the 11th Edition of the Fair-Rite catalog.

Summary: With no ferrite core present one has a low Q low inductance inductor. If one could construct a fully flux-coupled ferrite 61 core (LFEF~1.0), the Q at 1 MHz would be 1/ui*CLF=267. Highest Q occurs with an optimum value of the conductor at the high end of the band more than at the low end. The use of litz wire reduces the loss across the band, but more so at the high end. If solid wire is used, spaced-turns winding will reduce the losses from proximity effects. An advantage of using Litz wire in a ferrite-rod inductor is that there seems to be little downside to Q from close spaced winding. This helps with obtaining a larger inductance with a smaller solenoid and ferrite core. The use of larger diameter wire to reduce one of these losses usually has the effect of requiring a larger solenoid and ferrite core in order to keep the inductance the same, requiring mind-numbing tradeoffs. Experimentation with 4" long by 1/2" diameter ferrite 61 rods and litz wire of 50/46, 125/46, 270/46 and 420/46 construction with an inductance 250 uH suggest that a winding length of about 1.5" of close-wound 125/46 litz wire is close to optimum, from the standpoint of Q. I've tried to use 660/46 litz with a 4"x1/2" ferrite 61 rod to attain a high Q inductance of about 250 uH. It never worked, probably because the length of the rod, being close to that of the solenoid, caused a high flux density condition to occur near the ends of the solenoid, creating extra copper loss. Lesson: For high Q, the rod should be longer than the solenoid, maybe three times as long.

Ferrite cores of the same specification often exhibit rather wide variations in their ferrite loss-factor (thus affecting the attainable Q when used as a core). They also vary, to a lesser degree, in initial permeability (µi). This affects the inductance. Generally, when selecting cores from a group having identical specifications, the ones with the least initial permeability will have the least hysteresis loss, especially at high frequencies. This provides a convenient way to select cores that will yield the highest Q coils, without actually measuring Q: Wind a solenoid on a thin walled, low loss form and measure its inductance after placing each core, in succession, centered in

When using the "Selective" antenna connection, the antenna-ground components are connected through Cc across only 1/3 of the tank coil turns. This results in a reduction to about 1/9 of the resistive loading by the antenna on the tank, compared to the loading in the "Broad" connection. See Fig. 14. This reduced loading increases the loaded circuit Q, and hence selectivity. The ratio of unloaded to loaded Q is reduced, thus reducing sensitivity.

For practical purposes the 'leakage inductance' between that part of the primary that is bifilar wound with the secondary is very low. To the extent that it is not zero, it and the leakage inductance between the outer turns of the primary and the inner bifilar-ed 25 turns can be considered to be an added "leakage inductance" in series with C2 in Fig. 7; and in series with C3 in Fig. 14. The main effect of this leakage inductance, compared to having none, is to somewhat lower the highest frequency than can be tuned. The low end of the tuning range will be extended a small amount.

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ARTICLE 20

How to measure the impedance of an AM-band antenna-ground system, what one can do with the results, along with some measurements

Quick Summary: This Article describes a method to measure the series capacitive and resistive parameters of the impedance of an antenna-ground system vs frequency. Results from measurements on an attic antenna are given.

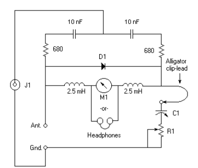

Fig. 1 - Ver. A of Antenna-ground measurement bridge.

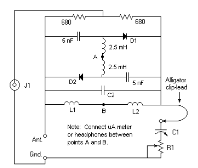

Fig. 2 - Ver. B of Antenna-ground measurement bridge.

Schematic using a half wave rectifier Schematic using a full wave rectifier

The circuit in Fig. 1 was inspired by an Article in The Crystal Set Society Newsletter of Jan 1, 1995. It was written by Edward Richley. He used a 1 MHz crystal oscillator for his source, so had no problem with using a 200 uA meter. I use a sine wave function generator for my RF source, but a radio Service man's oscillator may also be used if it has enough output. Either of these sources cannot supply as much signal as the xtal oscillator, so I had to increase sensitivity. That's what the 2.5 mH chokes and 5 nF caps are for. The 2.5 mH

creates the effect of adding a parallel RL in series with the air coil. When a core is inserted into the air-cored solenoid, the series resistance of the solenoid in air is increased by the factor (FDFa-1) to account for the increased power loss in the copper wire caused by the increased flux density from the core

Magnetic flux density surrounding the solenoid turns conductor is not uniform along the length of the solenoid. It is greater at the ends than along its central part. Increasing the length/diameter of the solenoid reduces the percentage of total flux that penetrates the copper and thus reduces resistive copper losses (especially at the two ends of the winding). Increasing the ratio of the length of the ferrite rod to that of the solenoid further reduces the percentage of total flux that penetrates the copper, further reducing resistive losses.

The amount of electric field that penetrates the core is important, especially at the high end of the band . The Nickel/Zinc cores such as type 61 have a very high resistivity dielectric as well as a rather low dielectric constant (ε) that has a high dielectric loss tangent (tan δ): Losses caused by the high (tan δ) may be minimized by using construction methods that keep the parts of the solenoid that are at a high electrical potential spaced away from the core. For instance a coil former sleeve made of low loss, low dielectric constant material can be used to isolate the high impedance parts of the solenoid from the core.

I-squared-R resistive power loss in the conductor caused by the AC current flow: Increased series resistance of the solenoid reduces Q, especially at the low end of the band compared to the high end, since the inductive reactance is at a minimum there (if the resistance, as a function of frequency is constant). Proximity and skin effect losses increase the RF resistance of

La=inductance of solenoid in air
FDF(suffix a or f)=flux density factor. This is a number, greater than 1, that corrects the value of the series DC resistance of the solenoid to its actual AC value. The "a" denotes air surrounding the solenoid, the "f" denotes the inclusion of a ferrite core.
LFEF=leakage flux effect factor, the ratio of Lp to µi*La. (LFEF is always less than 1)
when the solenoid conductors are bathed in magnetic flux. It has a suffix"a" when referring to the solenoid in air, and a suffix"f" when referring to a solenoid having a ferrite core.
Lp=parallel inductance representing the increase of inductance caused by the ferrite core. Lp=LFEF*ui*La Henrys
Ra=series RF resistance of solenoid in air
Ra*(FDF-1)=additional series resistance caused by increased average flux density around the conductors when a ferrite core is placed inside the solenoid
Rdc=DC resistance of the wire used in the solenoid
Ro=represents resistive power loss in Co. This loss in Co is contributed by the dielectric loss tangent of the ferrite and that of the solenoid former, if one is used.
Rp=parallel RF resistance across Lp, representing magnetic losses in the ferrite core. Rp=LFEF*ω*La/CLF Ω
ui=initial permeability of the ferrite. [125 for type 61 ferrite]
Qa=Q of the solenoid in air Qa=ω*La/(Rdc*FDFa)
Qt=Q of the real-word ferrite-cored inductor as represented in Fig. 1
ω=2*pi*frequency

The simplified equivalent circuit shown in Fig. 1 provides a convenient way of think about the effect of placing a ferrite core in a solenoid. Ra and La represent the resistive loss and inductance of the air-cored solenoid without the core (we want to increase the resultant Q and inductance). Adding a core

chokes eliminate RF loading by any resistive component of the meter or phones on the diode detector. The 5 nF caps eliminate resistive DC loading on the detector from the two 680 ohm resistors. I lay out the components breadboard style on a nonconductive table to minimize stray capacity, keep connections short, and especially keep the signal source lead of J1 away from the connections to each end of D1. In my setup D1 is a 1N34A, M1 is a 0-20 uA DC meter, R1 is a 75 ohm non-inductive carbon pot and C1 is a two gang variable cap of 365 pF per section. I parallel the two sections when the antenna capacitance is above 365 pF. A lower sensitivity meter can be used than the one used here, at the cost of a requiring a higher applied signal to J1.

If a sensitive enough meter is not available, a pair of high impedance phones (2000 ohms DC resistance) or preferably, a sound powered pair with the elements wired in series can be used. In this case, the generator must have its AM audio modulation turned on at its highest level. A modulation frequency of about 1 kHz is recommended. If the meter is used, do not connect the phones. If phones are used, do not connect a meter.

To use the bridge, tune the generator to a frequency of interest. Adjust C1 and R1 for minimum deflection on M1 or a null of the modulation tone in the phones. Increase the RF signal to J1 as much as possible in order to get the sharpest and most precise null. Measure the resistance of R1 with an ohmmeter. Use any desired method to measure C1. I use the cap. measurement range of my Fluke DVM. I'm sure the reader does not need to be reminded that this test involves radiating a weak RF signal from the antenna when making the measurements, so the length of time the generator is on should be kept as short as possible.

Possible issues: More sensitivity is needed or interference from antenna pickup of local stations obscures the bridge null.

If insufficient signal is available from the RF generator to provide satisfactory meter readings, one can use the more sensitive broadband circuit shown in Fig. 2. The values of L1 and L2 are 2.5 mH and C2 is set to zero in the broadband version. A full wave rectifier is used instead of the half wave one used in Fig.1 and it gives about twice the output. One can also change from using 1N34A diodes and try Schottky Zero Bias detector diodes such as the Agilent HSMS-2850 in either circuit. The HSMS-2855 Zero Bias diode is especially suitable for use in the circuit shown in Fig. 2 since it is a package having two independent diodes, one for D1 and the other for D2. One must be cautious when using the HSMS-2855 because the diodes can be damaged by the application of too strong a signal to J1. This can happen if the signal generator signal is very strong when the bridge is greatly unbalanced. It's best to start with a weak signal, balance the bridge, then increase the signal if necessary.

If the signal from the RF generator is not strong enough to override local pickup, thus obscuring the meter null, selectivity may be added to the bridge shown in Fig. 2 by making use of C2 and changing L1 and L2. If L1 and L2 are changed to, say, 10 uH inductors and C2 is made equal to 1200 pF, the bridge will be tuned to about 1 MHz. These changes will reduce the influence of local pickup upon measurement of antenna-ground impedance at 1 MHz. One suitable 10 uH inductor is Mouser's "Fastron" #434-23-100.

If one uses headphones instead of a meter as the null indicator, even greater sensitivity can be achieved by AF modulating the

* dielectric constant (ε), dielectric loss tangent (tan δ) and length of the 'former' upon which the solenoid is wound (if one is used)
* resistivity of the ferrite rod
* length (lf) and diameter (df) of the rod, and their ratio
* length (ls) and diameter (ds) of the solenoid, and their ratio
* Ratio of the length of the solenoid to the length of the ferrite rod
* size and type of wire (solid or litz) and spacing of the turns

A simplified equivalent circuit model for a ferrite cored inductor is shown in Fig. 1.

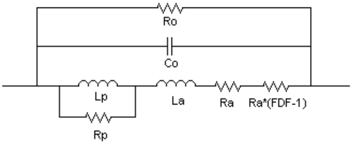

Fig. 1 - Simplified equivalent circuit of a ferrite cored inductor

Simplified equivalent circuit of a ferrite cored inductor

CLF=ferrite core loss-factor at a specific frequency [(30*10^-6) at 1 MHz for ferrite 61]
Co=distributed capacitance. This is made up of mostly capacitance from the hot parts of the solenoid, through the ferrite dielectric, to ground (assuming that one end of the solenoid is grounded). If a solenoid former is used, its dielectric is in the path and will affect the overall loss. Another part of Co is made up of capacitance from the hot parts of the solenoid through air, to ground.

ARTICLE 29

About Maximizing the Q of solenoid inductors that use ferrite rod cores, including charts of magnetic flux density and flux lines, with some actual Q and inductance measurements from simulations in FEMM

Summary: Many factors interact to affect the Q of ferrite rod cored inductors. Part one of this Article identifies and comments upon some of them. A simplified model and an equivalent circuit is also discussed. The second part describes several ferrite cored inductors, along with measurements of inductance and Q. The third part displays graphs of flux density, flux lines, inductance and Q of several ferrite cored inductors. The fourth part shows how flux density changes when an air gap is inserted in the center of a ferrite rod. It also includes simulations showing the distribution of magnetic flux density and current density in the turns of the solenoid. The fifth part shows how Q varies when the ratio of the length of the solenoid to the core changes Also shown is a chart showing the change of Q when the conductor spacing is changed. The sixth part discusses important info about ferrite 61 and similar materials.

Part 1: Modeling ferrite cored inductors

Bulk factors that affect inductor Q:

* Initial permeability of the ferrite material (μi) and ferrite loss-factor (LF)
* dielectric constant (ε) and dielectric loss tangent (tan δ) of the ferrite core

bridge signal generator and connecting a parallel L/C tuned to the modulating frequency of the generator across the phones. This will filter out much of the interfering cross talk from local pickup and pass the modulation tone with little loss. Suggested values are L=47 mH and C=0.5 uF if the modulating frequency used is about 1 kHz. A low cost coil having an inductance of 47 mH and a Q of about 9 at 1 kHz is available from Mouser as a Fastron Plugable Shielded coil, #434-02-473J ($1.20 each). Greater selectivity against cross-talk can be obtained by decreasing the inductance and increasing the capacitor.

I live about 9 miles from WOR and 12 from WABC, both 50 kW stations. 10 volts peak-to peak applied to the bridge overrides the local radio station pickup sufficiently to provide a clear null on the meter when using the circuit shown in Fig. 1 when using a 1N34A diode. A useable null with an applied signal of only 1.5 volts p-p can be obtained when using the circuit in Fig. 2 with zero bias detector diodes, sound-powered phones instead of a meter and the parallel LC filter.

Notes:

* If the RF source has too great a harmonic content, the bridge balance null will become less deep and sharp. That's why I used a sine wave function generator to assure a low harmonic content. If one uses a function generator for pure sine waves, make sure the symmetry control is set for best symmetry (minimum reading on the bridge microampmeter). In April 2004 Tom Polk published a description and schematic for a low distortion medium wave home brew signal generator. It looks very good, and can be found at: http://www.beecavewoods.com/testequipment/sinewave.html .

* If the resistance of a specific antenna-ground system is greater than 100 ohms, use a pot of a higher value than 100 ohms.
* A typical antenna-ground system will show a capacitance of a few hundred pF at the low end of the BC band. Because of the series inductance in the system, the measured capacitance will rise at higher frequencies. At a high enough frequency the system will go into series resonance and the bridge will not be able to be balanced. To measure the system series resistance at or above this resonance, place a hi Q capacitor of, say 100 to 220 pF in series with the antenna. That will raise the resonant frequency sufficiently so that the capacitor-antenna-ground circuit will be capacitive, a null can be obtained and the resistive component determined. An NPO ceramic or mica cap should be OK.
* At my location, detected signals from local strong stations show up as fluctuations at about 15% of full scale on the meter, but are not strong enough to obscure the bridge nulls from of the signal generator's signal.
* Unless the signal generator connected to J1 is battery powered (most aren't), it is important to put a common-mode radio frequency choke in the power line to the generator. I made mine by bundling a length of 18 ga. lamp cord into an 18 turn coil having a 9 inch diameter, and then fitting a male AC plug on one end and a female socket on the other. The turns were kept together using twist ties.

What can one do with the measurement results?

The main practical thing one can do with the bridge is to Measure and Monitor antenna-ground circuit resistance. This resistance comes primarily from the physical ground, not the antenna and ground connecting leads or radiation resistance of the antenna. Any increase in the antenna-ground resistance

#28 Originally published as Article #15 in Feb. 2001; later withdrawn. Republished as Article #28: 03/04/05

Practically speaking, what does all this mean? Mainly, improved theoretical understanding of diode detectors. (see last bullet point)

* Compared to the impedance matched condition, an increase in the volume of weak stations can be achieved if the RF source resistance driving the detector is dropped to 1/2 its RF input resistance, leaving the diode and output audio matching unchanged. Looking at it in another way, again compared to the impedance matched condition, the diode could be be replaced with one having half the Is, and the output load resistance doubled. Put a third way, the audio load resistance should be twice the RF source resistance and be equal to the axis-crossing resistance, Rd, of the diode.
* A higher value for the audio load resistance may be created by changing the impedance transformation of the audio transformer. Be aware that the insertion power loss of an audio transformer tends to increase when one operates it at higher input and output resistances than it was designed for. It is possible for increased transformer loss to cancel out some of the 2.0 dB improvement. See Article #5 for info on loss in various audio transformers. Fig. 4 and Part 4 provide info on some ways to audio-impedance match diodes having a high axis-crossing resistance (ones having a low Saturation Current).
* Changing to a diode having a lower Is will increase selectivity since the RF loading resistance value of the detector, on the tank is increased. This will increase the loaded Q of the detector tank, but will also increase the overall insertion power loss caused by the inherent losses in the tank inductor and tuning capacitor, nullifying some of the 2 dB improvement.
* By cut-and-try, many crystal set experimenters probably have already converged their designs to include this info.

serves to reduce the signal power available from the antenna. Any decrease, of course increases it. A halving of the antenna-ground system resistance provides a 3 dB increase in available signal power, if one properly rematches to the crystal radio set input circuit.

Measure: One can experiment with different grounds and various ground paralleling schemes to come up with the one that has the lowest resistance. Use of this one will result in maximizing the available signal power (more volume). Experiments using a counterpoise ground can be made.
Monitor: As has been recently been posted on the Yahoo Club: thecrystalsetradioclub, earth ground resistance deteriorates (increases) over time. This results in a gradual decrease in available signal power (less volume). Periodic measurement can alert one if this is happening so steps can be taken to correct the problem.

The other thing one can do, if one is mathematically engineering a crystal radio set, is to use the R and C values as parameters in the design. See Article #22.

Measurement results on an indoor attic antenna system:
My present external (as opposed to loop) antenna is in the attic. The horizontal element used to be made up of 7 twisted strands of #26 copper wire (17 ga.), suspended by strings about 1 1/2 feet below the peak of an asphalt shingled roof. It runs along under the peak and parallel to it for 53 feet. The wire is about 24 feet above ground level. The lead-in, connected to the center of the horizontal wire, runs horizontally, at a right angles for about 9 feet and then drops down vertically to the crystal radio set location, about four feet above ground level. The ground system consists of a connection to the cold water

supply in parallel with a connection to the hot water baseboard heating system. To achieve a low inductance ground connection I use 300 ohm TV twinlead, both conductors soldered in parallel, for each lead. The addition of a connection to the AC neutral does not seem to reduce the inductance or resistance of this antenna-ground system. I always suggest trying the addition of a connection to the AC neutral. Sometimes it helps.

The measured antenna-ground system capacitance was 295 pF at 0.5 MHz, 325 at 1.0 MHz, 410 at 1.5 MHz and and 660 at 2.0 MHz initially. The respective series resistances measured: 17, 12, 10 and 14 ohms. The equivalent reactance elements of this antenna are a capacitance of 285 pF in series with an inductance of 12.5 uH. Since my ground is composed of the house cold water supply pipes in parallel with the the hot water baseboard heating system pipes, much of the capacitance from the horizontal attic antenna wire is to them and the roof, not a real resistive earth ground. That, I think explains the low resistance and high capacitance readings. Probably the ground system is acting as a sort of counterpoise.

I decided to see if I could get greater signal pickup by changing to a very crude simulation of a flattop antenna. To do this, I paralleled the antenna wire with a piece of TV twinlead connected to it at each end and at the point of down-lead takeoff. The twinlead was separated from the 7/26 wire by about 2 1/2 feet. The new measured antenna-ground system parameters became: Capacitance: 430 pF at 0.5 MHz, 510 at 1.0 MHz and 860 at 1.5 MHz. The respective series resistance values became: 15, 12 and 11 ohms. The equivalent reactance elements became a capacitance of 405 pF in series with an inductance of 14.2 uH. Signal pickup increased by a

1. Series 1: The diode has an Is of 38 nA and an n of 1.03. R1= R2=Rd The graph shows that Ri is about 700k ohms at low input power levels and that it decreases towards 350k ohms at high input levels, where the detector acts as a peak detector. Not graphed, Ro, changes from about 700k to 1400k ohms as the input power goes from -95 dBW to -50 dBW. It is 1190k ohms at the LSLC input power point of -78.9 dBW.
2. Series 2: The Is of the diode is changed to 19 nA. All else stays the same. Ri approaches 1400k ohms at low input power levels and decreases towards 350k ohms at high levels. Not graphed, Ro is approximately 1400k ohms at low input power levels and 1400k ohms at high output levels.
3. Series 3: The diode Is stays at 19 nA, the n at 1.03, R1 at 700k ohms and R2 is changed to 1400k ohms. Ri changes from about 1400k to 700k ohms when going from low to high power levels. Not graphed, Ro is approximately 1400k ohms at low input power levels and 1400k ohms at high levels.

Look at Fig. 3. At the -95 dBW end of the graph, one can see that changing Is from 38 to 19 nA, and keeping R1 and R2 at at 700k ohms reduced the detector insertion power loss by about 1.5 dB (Series 1 to series 2). This comes from the increased voltage, Vi, at the detector input. Raising the output load resistance from 700k to 1400k ohms reduces the mismatch loss at the output to approximate. zero and reduces the overall insertion power loss by another approximate. 0.5 dB for a total improvement of 2.0 dB at low input power levels (Series 3). The detector insertion power loss at the -50 dBW input level is is also reduced by 0.5 dB because of the elimination of the output impedance mismatch.

the diode D with a different one having 1/2 the saturation current, Vi will increase. The reason is that the detector input resistance Ri is now twice R1. The voltage divider made up of R1 and Ri will reduce Vi to only two thirds of V1, making the new value of Vi=4/3 the old value. Since the detector is operating in square law mode, the internal source voltage in the detector that drives its output terminals will be (4/3)^2=1.77... times as much as before. This higher voltage will be divided down by the voltage divider action of the now twice-as-large diode output resistance and R2 to give an output voltage 1.185 as large as before. This equates to an output power 1.476 dB greater than in the original impedance matched condition.

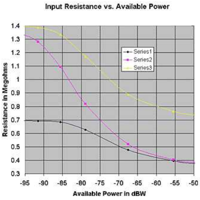

Fig. 2

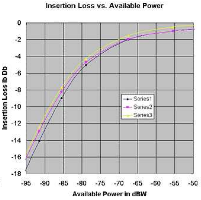

Fig. 3

Theoretical calculation and SPICE simulation show that that in a crystal set having equal values for R1 and R2, the diode parameters that give to lowest insertion power loss at low signal power levels fits the equation: R1=R2= 2*(0.0257*n/Is) at room temperature.

Now let us look at the effect on Ri and insertion power loss if Ri does not match R1. Look at Fig. 2.

negligible 0.8 dB at 710 kHz, and even that was, I'm sure, within experimental error.

One may want to compare these equivalent impedance components with the 'Standard Dummy Antenna', as specified in 1938 by the IRE (Institute of Radio Engineers) in 'Standards on Radio Receivers'. My reference for this is Terman's Radio Engineer's Handbook, first edition, 1943, pp 973 and 974. A rather complex equivalent circuit for the antenna is shown on page 974. It is stated that a simpler alternative network, given in footnote #2 on page 973, can be used when only the BC band is of interest. It consists of the series combination of a 200 pF capacitor, 25 ohm resistor and 20 uH inductor. Terman states that the two antenna equivalent circuits have closely the same impedance characteristics in the BC band. The impedance graph on page 974 and the impedance from the series combination of 200 pF, 25 ohm resistor and 20 uH differ, particularly in the resistive curve in the complex equivalent circuit. The 25 ohm resistance in the simplified circuit is probably taken from the resistance in the complex circuit, at the geometric center of the BC band. This resistance is shown as constant in the simplified circuit, and as a strong function of frequency in the more complex circuit. It is suggested that the complex equivalent circuit is theoretically derived, assuming a perfect ground and therefore does not include the resistance of the ground return path. The ground circuit can easily add 15-50 or more ohms to the circuit.

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ARTICLE 22

Design, construction and measurement of a single-tuned crystal radio set using a two-value inductor, along with a discussion of the cause of 'hash', short-wave ghost-signal and spurious FM reception. A way is presented for determining if the signal is operating a detector above or below its linear-to-square-law crossover point

Summary: This article describes Version 'b' of a single-tuned four-band crystal radio set, sometimes called a "Benodyne" (constant bandwidth with maximum weak-signal sensitivity across the whole BC band). It is an attempt to achieve the following two objectives at a -3 dB RF bandwidth of about 6 kHz (relatively independent of signal strength), and constant, high efficiency across the entire AM broadcast band by using two values of inductance in the tank: 1) Best possible sensitivity on weak signals; 2) Loudest possible volume on strong signals. Version 'c' of this crystal radio set uses Litz wire, introduces a contra-wound coil and has a "narrow selectivity" setting. It may be viewed in Article #26.

Means are provided for increasing selectivity at a small sacrifice in sensitivity. This crystal radio set is not designed to have strong immunity to local pick-up from local "blowtorch" stations. Selectivity and insertion power loss figures from a computer simulation are given and compared with those of the actual physical crystal radio set. A way to tell if the detector is operating below, at or above its 'Linear-to-Square Law Crossover point' (LSLCP) is described. No external antenna tuner is necessary. An explanation of 'short wave ghost signals' and 'hash' is provided along with some suggestions on how to combat them. This version 'b' uses only one diode and audio transformer configuration, as compared to the two used

approach the same value and equal 0.0257*n/Is = Rd ohms at room temperature when the input signal strength is low enough. See Article #0, part 4, Article 4, part 2 and Article 16 for information on Is and n, and ways of measuring them.

Fig. 1 represents a conventional diode detector. The tank circuit T is shown with no internal loss. A real world tank will have loss that can be represented by a shunt resistor connected across it. For convenience of analysis it is assumed that this loss resistance is absorbed into the source V1, R1 (For a more complete explanation, see Article #1, first paragraph after the third schematic.). Assume that the impedance of the source (antenna) and load (headphone) are transformed to equal values (R1 = R2) and select a diode that has an Rd equal to them. This will result in a reasonable impedance match at both the input and output if the signal power level places operation below +10 dB of the linear-to-square law crossover point. Little of the input power directed towards the detector will be reflected back to the source and most all of the output power from the detector will be absorbed in the load, R2. If the diode detector were a linear device with linear input and output resistances, this impedance-matched condition would result in the least detector power loss (greatest sensitivity) obtainable. It would seem clear that the crystal set detector could not be made more sensitive. Actually, not so! Very weak signal sensitivity can be improved by about 2 dB by appropriate mismatching at the input. This provides a rather hard-to-hear increase in volume, but every little bit helps.

In the impedance matched condition discussed so far, R1=Ri=Rd and R2= Ro. Simple math shows that the detector input voltage Vi will equal one half the internal source voltage V1. If we create an impedance mismatch between the source resistance R1 and the detector input resistance Ri by replacing

See Article #15a, Figs. 2 and 3 for info on the LSLC point of a diode detector.

Definition of terms:

Rd: Resistance of a diode at its axis crossing. Rd=0.0257*n/Is at 25 degrees Celsius
Ri: RF input resistance of the detector
Ro: Output resistance of the detector
R1: RF source resistance looking toward the tank
R2: Output load resistor
Is: Diode Saturation Current
n: Diode ideality factor
T: Parallel LC tuned circuit
LSLC point: The detector operational point halfway between linear and square law operation
Plsc(i) Max. available input power at the linear-square-law crossover point

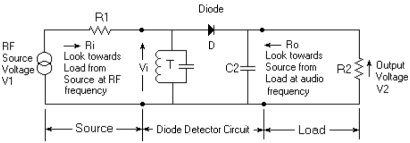

Fig. 1 - Diode Detector Schematic with Signal source and Output Load.

Diode Detector Schematic with Signal Source and Output Load

It has been asserted in these Articles that the RF input and audio output resistances, Ri and Ro, of a diode detector,

in the original obsolete version (now called Version 'a'). Also a new way to make a higher Q, low inductance coil using all the wire and coil form of the high inductance coil is described. Finally, the small performance sacrifice at the high end of the band that occurs when more readily available and lower cost parts are used is discussed.

Additional benefits of the "Benodyne" type of tank circuit are: (1) Reduction of the the sharp drop in tank Q or sensitivity at the high frequency end of the BC band often experienced when only one value of tank inductance is used for the whole BC band. (2) Reduction of the tank Q from loss in the variable cap when using lower cost units that use phenolic insulation, such as the common 365 pF cap (see Figs. 2, 3, 4, and 5 in Article #24). The "two inductance value benodyne* circuit is used in the crystal radio sets in Articles #22 and #26. We assume here that the two "Benodyne" component inductors (see "The Tank Inductor" in Article #26) provide a tank inductance of 250 uH in the low frequency half of the BC band (520-943 kHz) and 62.5 uH in the high half (0.943-1.71 MHz). If the large 250 uH inductance setting were used all the way up to the top end of the BC band (as in the usual case), a total tuning capacity of 34.7 pF would be required at 1.71 MHz (Condition A). In the "Benodyne" circuit, with the 62.5 uH inductance setting used for the high frequency half of the BC band, a total tuning capacitance of 139 pF is required at 1.71 MHz (Condition B). Benefit (1) occurs because in condition A, a larger fraction of the total tuning capacitance comes from the typically low Q distributed capacity of the inductor than in condition B. This results in a higher Q total capacitance in condition B than in condition A. Benefit (2) occurs because The effective Q of a typical 365 pF variable cap, when used with a 250 uH tank, is about 500 at 1.71 MHz (see Fig. 3 in Article #24). The Q of the 365 pF variable cap, when set to 139 pF, is greater than

1500 (see Fig.5 in Article #24). This higher Q results in less loss and greater selectivity at the high end of the BC band in condition 2. A further benefit of the Benodyne circuit at the high end of the BC band is greater immunity from the Q reducing effects of surrounding high loss dielectric materials such as baseboard etc. The lossy stray capacity introduced is better swamped out by the high shunt tuning capacity used.

The Crystal Radio Set Design, in a (large) Nutshell:

* The design approach is to divide the AM band into several sub-bands in an attempt to keep the selectivity constant and the insertion power loss low. Many concepts described in various Articles on my Web Index Page, as well as some new ones, are used in the design.
* The first step is to divide the band into two halves: Lo (520-943 kHz) and Hi (943-1710 kHz). Two-step shunt inductive tuning is employed to switch between bands. A tank inductance of 250 uH is used in the Lo band and of 62.5 in the high band.
* The Lo band is further subdivided into two sub-ands: LoLo (520-700 kHz) and HiLo (700-943 kHz). The Hi band is also subdivided into two sub-ands: LoHi (943-1270 kHz) and HiHi (1270-1710 kHz).
* Two different operational resistance levels at resonance, measured at the top of the tuned circuit (point 'A' in Fig. 5), are used at the center of the sub-bands. This impedance level is 125k ohms for the LoLo and LoHi bands and 250k for the HiLo and HiHi bands (excluding the resistive losses of the components used). These resistance values are made up of the parallel combination of the transformed RF antenna resistance and diode input RF resistance (at point A). These two resistances should be equal to each other to achieve a

ARTICLE 28

How to reduce diode detector weak signal insertion-power-loss to less than that possible when the input is impedance matched

Quick Summary: Diode detector insertion power loss can be reduced below the value achieved under impedance matched conditions provided it is operating below its LSLC point. The optimal conditions are: (1) The output audio load resistance equals twice the RF source resistance. (2) The Saturation Current and Ideality Factor of the diode are such that the very-low-signal output resistance of the detector (axis-crossing resistance, aka Rd, of the diode) equals the output load resistance. These conditions insure an impedance match at the audio output and a 1:2 mismatch at the RF input (source resistance = half the RF input resistance of the diode). Please bear in mind that the LSLC point is a point 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. Info on the LSLC point is available in Article #15a.

It has usually been assumed by myself and others that power loss in a two port device (here, a crystal set) is minimized when its input and output ports are impedance matched. This article will show that this is not true in the case of a diode detector operating at a signal power level well below that of its region of essentially linear operation. It is true, however, when a strong signal (well above the LSLC point) is being received. In the linear region, audio output power is proportional to RF input power. That is, for every dB of change in input power there will be one dB change of output power. In the lower power region, called the 'square law' region, a change of one dB in input power results in a two dB change in output power.

it simulate several in parallel. These approaches have not been experimentally investigated.

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minimum insertion power loss, at the design bandwidth. This means that the transformed antenna and diode RF resistances are each 250k in the LoLo and LoHi bands and 500k in the HiLo and HiHi bands at point 'A'. The two different transformed RF antenna resistance values (at point 'A'), at the top of the tank are achieved by proper adjustment of a variable capacitor in series with the antenna (C7 in Fig. 5). The two different diode RF tank loading resistance values (at point 'A') are achieved by tapping the diode onto the tank at a point that is 70% of the turns up from ground for bands HiLo and HiHi. The tank is not tapped for the LoLo and LoHi bands, and connection is to the top of the tank.

* The weak-signal audio output and RF input resistances of a diode detector are approximately the same and equal to 0.026*n/Is. The strong-signal audio output resistance of a diode detector approximately equals 2 times the RF resistance of its source. Compromise audio impedance transformation ratios are used to optimize performance on both weak and strong signals, thus maximizing sensitivity and volume.

* The design is scalable. Less expensive parts that may have somewhat greater losses may be used with some penalty in sensitivity and selectivity at the at the high end of the band. See the Parts List for a listing of more easily available and lower cost parts than the ones used in the original de sign. A tradeoff between sensitivity and selectivity can be achieved by changing the ratio of C7 and C8. Less capacitance in C7 increases selectivity and reduces sensitivity, and vice versa.

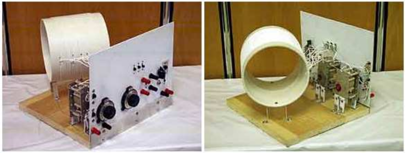


Fig. 1 - Single-Tuned Four-Band Crystal Radio Set, Version 'B'.

The design objectives for the crystal radio set are:

1. To achieve a relatively constant -3 dB bandwidth of 6 to 8 kHz across the full range of 520 to 1710 kHz, with a relatively constant RF power loss in the RF tuned circuits of less than 4 dB.
2. To to provide adjustment capability for greater selectivity, or less RF loss when needed.
3. To provide optimal performance with external antenna-ground systems having a fairly wide range of impedance.
4. To provide a simple-to-use switching setup for comparison of a 'test' diode with a 'standard one'.
5. To provide a volume control that has a minimal possible effect on tuning, having a range of 45 dB in 15 dB steps. This was incorporated in the design because the two local 50 kW blowtorch stations ABC and WOR (about 10 miles away) deliver a very uncomfortably loud output from SP headphones from my attic antenna. A means of volume reduction was needed. This method of volume reduction actually increases selectivity by isolating antenna-ground resistance from the tank circuit.
6. Introduce a new (to me) method for constructing high Q low inductance value inductors.

#11. The simulated crystal radio set consisted of two impedance-matching/tuning capacitors (C7 and C8 in Article #26) with a tank inductor having a Q value extrapolated from the values given in Table 4 of Article #26. The simulation program was 'SuperStar', by Eagleware. Those uncomfortable with the concept of 'Available Power' may find Part 3 in Article #0 helpful. Note that the diodes having the lowest n*Is value (and the lowest detector loss) result in the greatest loss from the tuning components in the crystal radio set. This means that to gain the greatest benefit from using a diode having a low n*Is, the parallel resonant loss resistance of the tank circuit must be made as high as possible(unfortunately reducing selectivity).

Note re diode performance when receiving strong signals: A high diode reverse-breakdown voltage rating is important in this case to prevent tank resistive loading from diode reverse conduction caused by the high reverse voltage present during the non-forward-conduction half-cycle. When this happens, volume is reduced. Diodes that have a high reverse breakdown voltage rating usually have a high value for the product of their saturation current and ideality factor and are best for obtaining maximum volume on strong stations. The diodes that are best for weak signal reception usually have a low n*Is product. Many of those who use the HP 5082-2835 report inferior volume on strong stations. I believe the cause is explained above. One solution to this problem is to provide switching means for two diodes as is done in the crystal set described in Article #26. Another possible approach might be to to use several HP 5082-2800 high-reverse-breakdown-voltage diodes in parallel. Several diodes would probably be needed because of the low saturation current of one '2800. One could also use only one '2800 and supply a little forward bias voltage to make digital voltmeter was connected to the "T" connector in Fig. 1 and used in the final measurements. Since there were internal noise issues with the Fluke, the 20 dB attenuator (SW3) in the AMCS was switched in to enable increasing the signal to the DVM by 20 dB to overcome the noise. The resistor values in the "inverted L" pad in the AMCS (45.0 and 5.55 ohms), along with the 25 ohm resistor, provide a source resistance of about 30 ohms and an attenuation of 22.98 dB, exclusive of any attenuation introduced by SW1, SW2 or SW3.

At first an HP model 3312A function generator was used as the RF source. Final measurements quoted were made using an HP model 33120A synthesized signal generator.

Measurements of the diodes having the lower values of Is were made at 892 kHz (Band A, sub-band 1 of the crystal radio set). Insufficient impedance transformation range was available to match diodes having the higher values of Is, so those were measured at 1205 kHz (Band B, sub-band 3). These frequencies were chosen so as to eliminate signal pickup from local stations.

The actual insertion power loss in the crystal radio set caused by losses in its L and C components was accounted for in each diode measurement by feeding an RF signal of -84.95+20.98+20+X dBW into the AMCS. (The raw internal power loss in the AMCS is 22.98 dB, and the 20 dB attenuator (SW3) was activated). X represents the L/C losses from the tank inductor and C7 and C8 in the crystal radio set used for the tests. Its value was determined from a computer simulation of the crystal radio set, using a source resistance of 30 ohms and a load resistance equal to the Rxc of the diode to be tested, and noting the insertion power loss as X. The value of X varied from 0.378 dB for diode #1 up to 1.711 dB for diode

1. Theory

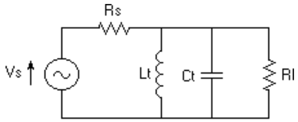


Fig.2 - Single-tuned circuit

Basic single tuned schematic

The frequency response shape of the circuit shown in Fig. 2 is that of a simple single tuned circuit and can be thought of as representative of the nominal response of a single tuned crystal radio set. Consider these facts:

1. If Lt and Ct have no loss (infinite Q), zero insertion power loss occurs at resonance when Rs equals Rl. This is called the 'impedance matched' case. The power source (Vs, RS) sees a resistance value equal to itself (Rl). Also, the load (Rl), looking towards the input sees a resistance (RS), equal to itself. In the practical case there is a finite loss in Lt and CT This can be represented by an additional resistance Rt (not shown), shunted across the tuned circuit. The input resistance seen by (Vs, RS) is now the parallel combo of Rt and Rl and it is less than RS The impedance match seen by (Vs, RS) when the tank Q (Qt) was infinite is now destroyed. The impedance matched condition can be restored by placing an impedance transformation device between the source, (Vs, RS) and the tank. In the crystal radio set to be described, the Q of the highest Q practical inductor thought suitable for the design was found to be sufficient to enable about a 6 kHz loaded

bandwidth to be obtained with about a 4 dB insertion power loss over the tuning range of 520-1710 kHz.

2. In Fig. 2, if tuning could be done with Lt alone, leaving CT fixed, the bandwidth would be constant. The problem here is that high Q variable inductors that can be varied over an approximately 11:1 range, as would be needed to tune from 520 to 1710 kHz do not exist. On the other hand, tuning by varying the value of CT by 11:1 will cover the range, but have two disadvantages. (1) Bandwidth will vary by 1:11 from 520 to 1740 kHz. (2) In the practical case, if the bandwidth is set to 6 kHz at the low end of the band, and an attempt is made to narrow the bandwidth at 1710 kHz by placing a capacitor in series with the antenna, the insertion power loss will become great.

3. The compromise used here is a coil design that can be switched between two inductance values differing by 4:1. The high inductance setting is used for the low half of the band and the low inductance setting for the high half. Capacitive tuning is used to tune across each band. The technique used here, in creating the two inductances, enables the Q of the low value inductance to to be much higher than would be the case if a single coil of the same diameter but with fewer turns was used. This technique uses two coils, closely coupled, and on the same axis. They are connected in series for the large inductance and in parallel for the small one. The small inductance has a value 1/4 that of the large one and about the same Q at 1 MHz. The innovation, so far as I know, is to use the full length of wire used in the high inductance coil, occupy the same cubic volume, but get 1/4 the inductance and keep the same Q as the high inductance co il.

4. The high and low bands are each further subdivided giving a total of four bands (LoLo, HiLo, LoHi and HiHi). If this was not done, we would be faced with a bandwidth variation of about 1:3.3 in each band. The bands are geometrically divided

Two charts are presented in Article #16 showing measurements of Is and n for 10 different diodes. The Schottky diodes seem to have fairly constant values of Is and n as a function of current. The silicon p-n junction and germanium measurements show how Is and n can vary, in other diode types, as a function of current.

Appendix: The objective of these measurements is to measure the performance of various diodes when used as detectors; at a signal level well below their LSLCP so that their weak signal performance can be compared. The measurements described in this Article were made with an 'available RF power' of -84.95 dBW (3.2 nW) applied to the diode. Here is how that value was chosen:

Initial measurements were made using a Tektronix model T922 scope having a maximum sensitivity of 2 mV/cm. The scope was connected to P1 in Fig. 1; the horizontal sweep rate was set to display about 3 cycles of RF. The RF voltage that could be read on the scope, with reasonable precision, was considered to be about 2 mV minimum peak-to-peak, providing a vertical display of 1 cm. The voltage at P1 drives a 25 ohm resistor, the resistance of which is transformed in the crystal radio set up to a value that matches the input resistance of the diode. The correct input impedance match is attained by interactively adjusting C7 and C8 on the crystal radio set to maximize the rectified DC output voltage. 2 mV pp RF voltage equates to 2/(sqrt8) mV RMS. Since available power=Pa=(Erms^2)/(4*source resistance), Pa=5 nW. If one allows for about 2 dB loss between the input to the crystal radio set and that to the diode, the available power that actually reaches the diode becomes about 3.2 nW. To obtain better measurement precision a Fluke model 8920A true RMS RF audio distortion will come in sooner on strong signals, because of its low reverse breakdown volt age. It performs best if used in a crystal radio set having RF source and audio load resistances of about 400k ohms, rather high values. An excellent diode for both weak and strong signal reception is the obsolete ITT FO-215 germanium diode, still available, think from Dave Schmarder at http://www.1n34a.com/catalog/index.htm . Crystal radio sets having RF source and audio load resistances of about 200k ohms may have better sensitivity with two of the HSMS-286L arrays in parallel. One section of the Infineon BAT62-08S triple diode should work the same as two HSMS-286L arrays in parallel. Agilent semiconductors are carried by Newark Electronics and Arrow Electronics, among others. Agilent or Infineon may sometimes send free samples to experimenters who ask for them.

The apparent error in output power for diode #1 has been checked many times. The figure appears to be accurate. I don't know the reason for the anomaly, except that the diode probably has an increased value for n*Is at the 21 Na rectified current, compared to the value of Is*n from the measurements in Table 1 (made at a higher current). It is known that there are extra causes for conduction in a diode beyond those modeled by the Shockley equation. Other measurements show that this diode also does not follow the Shockley diode equation at high currents (see Article #16). Diodes A, B and C are randomly selected 1N56 germanium units. They also show poorer performance than would be expected from Schottky diodes of the same Is and n. Note that germanium diodes diodes #1, A, B and C provide less output than would be expected from Equation #5. The output from germanium diode #5 is close to that expected from Equation #5 as is the output from all the Schottky diodes.

and are: 520-700 (LoLo), 700-943 (HiLo), 943-1270 (LoHi) and 1270-1710 (HiHi) kHz. The bandwidth should vary 1:1.8 across each sub-band. The same relation should hold between the HiHi and LoHi band. The bandwidth at the center of each of the four bands are made approximately equal to each other by raising the loading resistance of the antenna and diode on the tank by a factor of two in band HiLo compared to the value used in band LoLo. The same adjustment is used for bands HiHi and LoHi.

2. Design Approach for the Center of each of the four Bands.

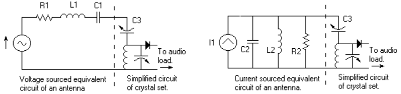


Fig. 3a Fig. 3b

Simplified Schematic of Crystal Set

Fig. 3a shows the simplified Standard Dummy Antenna circuit, described in Terman's Radio Engineer's Handbook, for simulating a typical open-wire outdoor antenna-ground system in the AM band. R1=25 ohms, C1=200 pF and L1=20 uH. See Article #20 for info on how to measure the resistance and capacitance of an antenna-ground system. The values shown for Fig. 3a are used in the design of the crystal radio set. R1 represents primarily the ground system resistance. C1 represents the capacitance of the horizontal wire and lead-in to ground and L1 represents the series inductance of the antenna-ground system.

The values of R1, C1 and L1 in Fig. 3a will be considered to be independent of frequency. To the extent that they do vary with frequency, C7 and C8 in Fig. 5 can be adjusted to compensate. The current-source equivalent circuit of the antenna-ground circuit is shown in Fig. 3b. To a first degree of approximation, in the practical case, C2 in Fig. 3b is independent of frequency. R2 will vary approximately inversely with frequency. We will ignore the effect of L2, since its value is large, except when approaching the first resonance of the antenna-ground system. The design approach is to place a variable capacitor C3 in series with the antenna circuit (Fig. 3a) to enable impedance transformation of the antenna-ground circuit to an equivalent parallel RC (Fig. 3b), the R component of which can be adjusted by changing the value of the C3 to follow a desired relationship vs frequency. One of the objectives of the design is to enable as constant a bandwidth as possible vs. frequency. This requires the aforementioned equivalent parallel R2 component to vary proportionally with the square of the frequency if capacitive tuning is used as it is here, in each sub-band (Q must be proportional to frequency for a constant bandwidth). This design attempts to accomplish this in the center of each frequency band. Performance is close at band edges.

3. The single tuned crystal radio set

The topology of the single tuned circuit is changed from band to band as shown in Fig. 4 below.

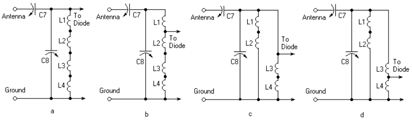


Fig. 4 - (a) Circuit topology for the LoLo band. (b) Circuit topology for the HiLo band. (c) Circuit topology for the LoHi band. (d) Circuit topology for the HiHi band.

low loss manner. A low loss high input resistive source is easier to achieve with a high Q loop driven crystal radio set that uses the loop as the tank than with one driven by an external antenna and ground that uses a separate high Q tank coil. This is because one of the sources of loaded tank resistive loss, the external antenna-ground system resistance, is eliminated. The radiation resistance of the loop is usually negligibly small compared to the loss in the loop when considered as a stand alone inductor. It is assumed in this discussion that the diode is connected to the top of the loop, the point of highest source impedance. Don't take this as a recommendation to go to a loop antenna for the best weak signal reception. A good outside antenna-ground system will outperform a loop by picking up more signal power. Conclusion: A diode with the lowest n*Is may be theoretically the best, but achieving impedance matching of input and output may not be possible. In practice, a compromise must be struck between a diode with the lowest n*Is and one having a lower axis-crossing resistance (Rxc). This means, in general, a higher Is. It can be achieved by paralleling several diodes or using a different diode type.

A good diode array to try in high performance crystal radio sets intended for weak signal reception is an Agilent HSMS-286L, with all three diodes connected in parallel. This diode array is packaged in a small SOT-363 SMD package but is easy to use even without a surfboard to aid in its connection. The three anode leads exit from one side of the package with the three cathode leads from the other. A quick connection solder blobbing all three anodes together and to a thin wire, and a similar connection to the cathodes is easy to do. Use a low temperature soldering iron and as little heat as possible to avoid injuring the diodes. This triple diode performs about the same as six Agilent 5082-2835 diodes in parallel, except that

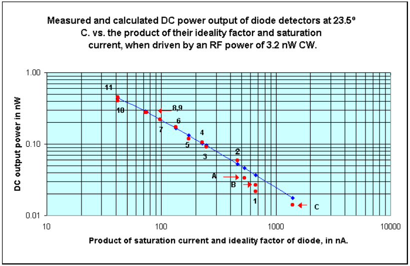


Fig.2

The red data points indicate actual measurements. The blue values are calculated from equation #5 in Article #15A. The blue line is a connection of the points calculated from equation #5. Note the close fit.

Discussion: Fig. 2 shows the close correlation of measured output power with measured diode n*Is, as predicted by equation #5 in Article #15A. This suggests that n*Is is a valid 'figure of merit' for a diode used to detect weak signals. Remember: The detector power loss figures shown in the last column of Table 2 would be even larger if the test signal of 3.2 nW were smaller. The assumption in all of this is that both the input and output ports of the crystal radio set are reasonably well impedance matched.

Note that the input and output resistances of a diode detector using diodes #10 and 11 are very high. Matched input source and output load resistances this high are hard to achieve in a

Four band equiv. ckt. schematic

The resonant RF resistance values at the top of C8 (Fig. 4), from the transformed antenna resistance, (at the center of each sub-band) are designed to be: 250k ohms for bands LoLo and LoHi, and 500k ohms for bands HiLo and HiHi. Since the diode is tapped at the 0.7 voltage point for bands HiLo and HiHi, it sees a source resistance at resonance of: 125k for bands LoLo and LoHi and of 250k ohms for bands HiLo and HiHi. These figures apply for the theoretical case of zero loss in the tuned circuits (infinite Q). In a shunt capacitively tuned crystal radio set, loaded with a constant resistive load, the bandwidth will vary as the square of the frequency. To understand why, consider this: When the resonant frequency of a tuned circuit loaded by fixed parallel resistance is increased (from reducing the total circuit tuning capacitance), the shunt reactance rises proportionally, giving rise to a proportionally lower circuit Q. But, a proportionally higher Q is needed if the bandwidth is to be kept constant. There for, the square relation.

In the practical case, we are faced with two problems. (1) How should we deal with the fact we work with finite Q components? (2) At high signal levels (above the LSLCP), the RF load presented by the diode to the tuned circuit is about 1/2 the audio load resistance, and at low signal levels (below the LSLCP) the RF load presented to the diode is about 0.026*n/Is ohms. Compromises are called for.

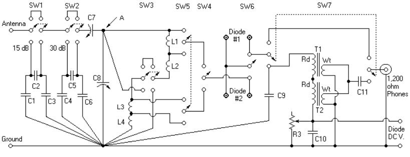


Fig. 5 - Schematic of Single-Tuned Four-Band Crystal Set (Ver. B).

Schematic diagram og version B

Parts List - All components are chosen for the best possible sensitivity at a -3 dB RF bandwidth of 6 kHz (except for not using litz wire in the inductor).

* C1, C3: 200 pF NPO ceramic caps.
* C2: 100 pF NPO ceramic cap.
* C4, C6: 270 pF ceramic caps.
* C5: 18 pF NPO ceramic cap.
* ** C7, C8: 12-475 pF single section variable capacitors, such as those that were mfg. by Radio Condenser Corp. They use ceramic stator insulators and the plates are silver plated. Purchased from Fair Radio Sales Co. as part # C123/URM25. Other capacitors may be used, but some of those with phenolic stator insulators probably will cause some reduction of tank Q. The variable capacitors are fitted with 8:1 ratio vernier dials calibrated 0-100. These are available from Ocean State Electronics as well as others. An insulating shaft coupler is used on C7 to eliminate hand-capacity effects. It is essential,

* Is and n were measured using forward voltages of 39 and 55 mV (average current of between 3.8 to 5 times Is).

Table 2. CW measurements of output power at 23.5° C.; detector input power: 3.2 nW (-85 dBW)

Diode #	Diode load in ohms*	Measured DC output in mV	Measured output in nW	Measured output in dBW	Product of n and Is in nA	Calculated output in dBW	Measured minus calculated output in dB	Detector power loss in dB!
C	29.5k	0.69	0.01614	-108.52	1391	-107.52	-1.00	**23.6**
1	47k	1.01	0.02160	-106.65	663.3	-104.50	-2.15	**21.7**
B	50.5k	1.25	0.03094	-105.70	665.5	-104.38	-1.32	**20.8**
A	63.0k	1.56	0.03875	-104.72	529.3	-103.41	-1.31	**19.8**
2	78k	2.15	0.05855	-102.33	459.3	-102.80	+0.47	**17.4**
3	145k	3.62	0.09038	-100.44	245.5	-100.63	+0.19	**15.5**
4	127k	3.65	0.1049	-99.79	228.3	-99.90	+0.11	**14.8**
5	191k	4.75	0.1181	-99.28	174.0	-98.80	-0.48	**14.3**
6	214k	6.11	0.1744	-97.58	133.2	-97.88	+0.30	**12.6**
7	296k	8.06	0.2195	-96.59	97.07	-96.40	-0.19	**11.6**
8	400k	10.48	0.2753	-95.60	74.0	-95.90	+0.30	**10.7**
9	400k	10.48	0.2746	-95.61	72.4	-95.44	-0.17	**10.7**
10	658k	16.3	0.4038	-93.94	41.8	-93.54	-0.40	**9.00**
11	676k	17.5	0.4530	-93.44	41.6	-93.55	+0.11	**8.5**

* Diode load is equal to Rx, its axis-crossing resistance

Note: The rectified DC current ranges between 21 and 29 nA, with most diodes close to 25 nA

Graph of Power Output vs n*Is

The input conditioning device (AMCS), used to aid in measurement of input power is shown in Fig. 2 of Article #11. There, it was used in a procedure to measure input AM sideband and output audio power. Here it is used as a convenient way to provide an accurate source voltage having a known internal resistance. A CW signal generator tuned to, say, 1 MHz is connected to the AMCS as the source of RF power. If the generator has an AM modulation capability, that can be used with headphones as an aid in initially tuning the crystal radio set to the test signal. Table 2 and Fig. 2 show the results of the measurements.

Table 1. Measured values of saturation current and ideality factor for some diodes, normalized to 25° C

Diode	Diode type	Is* in nA	n*
A	1N56 germanium marked GE	553	1.06
B	1N56 germanium marked GE	692	1.07
C	1N56 germanium marked GE	1317	1.17
1	Blue Radio Shack 1N34A germanium, no markings	678	1.09
2	Two high Is Agilent HBAT5400 Schottkys in parallel	438	1.16
3	Agilent high Is HBAT5400 Schottky	236	1.16
4	Infineon BAT62-03W Schottky	243	1.04
5	Radio Shack 1N34A germanium marked 12010-3PT	167	1.16
6	Infineon BAT62-08S Schottky	143	1.04
6.5	ITT FO-215 glass germanium (rare, although Mike Peebles and Dave Schmarder have them)	109	1.02
7	Agilent low Is HBAT5400 Schottky	104	1.04
8	Agilent HSMS-286L Schottky, all three diodes in parallel	78	1.05
9	Six Agilent 5082-2835 Schottkys in parallel	77	1.04
10	Agilent HSMS-282N Schottky, all four diodes in parallel	45.6	1.02
11	Four Agilent 5082-2835 Schottkys in parallel	44.9	1.03

for maximum sensitivity, to mount C7 in such a way that stray capacity from its stator to ground is minimized. See Part 9 for info on mounting C7. The variable capacitors used in this design may not be available now. Most any other capacitor with silver plated plates and ceramic insulation should do well.

*

C9: 47 pF ceramic cap.

*

C10: 0.1 to 0.22 uF cap.

*

C11: Approx. 1.0 uF non-polarized cap. This is a good value when using RCA, Western Electric or U. S. Instruments sound powered phones, with their 600 ohm elements connected in series. The best value should be determined by experiment. If 300 ohm sound powered phones having their 600 ohm elements connected in parallel are used, C11 should be about 4 uF and a different transformer configuration should be used.

*

** L1, L2, L3 and L4: Close coupled inductors wound with uniformly spaced Teflon insulated 18 ga. silver plated solid wire. This wire is used only to gain the benefit of the 0.010" thick low-loss insulation that assures no wandering turns can become 'close-spaced'. L1 has 12 turns, L2 has 8 turns, L3 has 6 turns and L4 has 14 turns. The coil form is made of high-impact styrene. I used part #S40160 from Genova Products (http://genovaproducts.com/factory.htm). See Fig. 6 for hole drilling dimensions.

*

** SW1, 2: DPDT general purpose slide switches.

*

**SW3, 4, 5 and 6: Switchcraft #56206L1 DPDT mini Slide switches. This switch has unusually low contact resistance and dielectric loss, but is expensive. Other slide switches can be used, but may cause some small reduction of tank Q. SW6 is used as a SPDP switch. Don't wire the two halves in parallel.

*

T1, T2: Calrad #45-700 audio transformer. Available from Ocean State Electronics, as well as others. If 300 ohm phones are to be used, see the third paragraph after Table 1.

*

R3: 1 Meg Pot. (preferably having a log taper).

*

Baseboard: 12" wide x 11 1/8 " deep x 3/4" thick.

*

Front panel is made of 0.1" high-impact styrene. Other materials can be used. I was looking for the lowest loss, practical material I could obtain.

** For lower cost, the following component substitutions may be made: Together they cause a small reduction in performance at the high end of the band (about 1.75 dB greater insertion power loss and 1.5 kHz greater -3 dB bandwidth). The performance reduction is less at lower frequencies.

* Mini air-variable 365 pF caps sold by many distributors such as The Crystal Set Society and Antique Electronic Supply may be used in place of the ones specified for C7 and C8.

* 18 ga. (0.040" diameter) "bell wire" supplied by many distributors such as Home Depot, Lowe's and Sears may be used in place of the teflon insulated wire specified. This vinyl insulated bare copper wire is sold in New Jersey in double or triple twisted strand form for 8 and 10 cents per foot, respectively. The cost comes out as low as 3 1/3 cents per foot for one strand. The main catch is that one has to untangle and straighten the wires before using them. I have used only the white colored wire but I suppose the colored strands will work

A 3.2nW (-84.95 dBW) un-modulated source of available RF power of is applied to the diode for all measurements. This power level is about 12 dB below the LSLCP(i) of the average diode used in these tests. An AM broadcast signal of this power level will result in quite a weak sound in SP phones impedance matched to the output of a crystal radio set. The RF input power is applied through the AMCS described in Article 11. If the input to it is monitored by a DVM connected to the "T" connector, the AMCS should be considered to be an attenuator having an input resistance of 50 and an output resistance of 30.244 ohms. Its attenuation is 22.975 dB. The crystal radio set used in these measurements is described in Article #26. It was used because its performance has been well characterized and its input impedance can be changed over a wide range. The output resistor into which the output power is dissipated is R3. The primary windings of T1 and T2 are shorted when taking measurements.

The measurement procedure, for each diode, consists of applying an RF power source having an available power of 3.2 nW to the diode detector, adjusting the input impedance transformation (C7 and C8 in the crystal radio set described in Article #27) for maximum output voltage and recording that value.

Block diagram

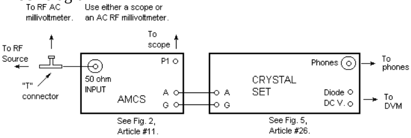


Fig. 1 - Connection block diagram.

ARTICLE 27

Measurement of the sensitivity of a crystal radio set when tuned to a weak fixed signal, as a function of the parameters of the detector diode; including output measurements on 15 diodes

Summary: This Article shows how the 'weak signal' power loss, from the resultant operation in the square-law region of the detector, of a crystal radio set, varies as a function of the parameters of the diode. The measurements clearly shown how 'weak signal' detector loss is reduced when diodes having lower values of the product of saturation current and ideality factor are used. This results in obtaining greater volume from weak signals. Actual measurements compare closely to those predicted from equation #5, developed in Article #15A. It is assumed that the input and output impedances of the detector are reasonably well matched.

Acronyms and Definitions of Terms

AMCS	Apparatus used when Measuring Crystal Radio Set Insertion Power Loss and Selectivity.
CRYSTAL RADIO SET	A crystal radio set, such as that described in Article #26 that has the capability of a continuously adjustable input impedance transformation.
Is	Diode Saturation Current in Amps
LSLCP(i)	Linear-to-Square Law Crossover Point in dBW (referred to the input power)
n	Diode Ideality Factor
Rxc	Axis-crossing resistance of a diode. Rxc=0.026*n/Is.

1. The Measurements:

The measurements of output power are made using a simpler and quicker method than that used in Article #11, since a CW instead of a modulated signal is used. This method involves measuring DC output voltage into a resistive load when the input of the detector is fed from a fixed source of available RF power.

the same (re dielectric loss). The measured outside diameter of the wire from various dealers varied from 0.065 to 0.079". The high dielectric loss factor of the vinyl, compared to the teflon specified above will cause some reduction of sensitivity and selectivity, more at the high end of the band than the low end. I don't think the difference would be noticeable to a listener.

* Radio Shack mini DPDT switches from the 275-327B assortment or standard sized Switchcraft 46206LR switches work fine in place of the specified Switchcraft 56206L1 and cost much less. See Article #24 for comparison with other switches. Any switch with over 4 Megohms Rp shown in Part 2 of Article #24 should work well as far as loss is concerned. Overall, losses in the switches have only a very small effect on overall performance.

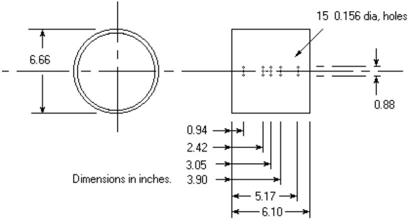


Fig. 6 - Coil form drilling dimensions (Version B).

Coil form drilling dimensions

The coil should be mounted with its axis at a 30 degree angle to the front panel as shown in Fig.1. The center of the coil form is 2 7/8" back from the rear edge of the baseboard and 5

5/8" to the left of its right edge. These dimensions are important, as is the actual size of the breadboard, if one wishes to construct a double-tuned four band crystal radio set out of two Version 'b' single-tuned four band crystal radio sets as described in Article #23.

Table 1 - Switch Functions for Version 'b':

SW1	15 dB volume control "capacitive" attenuator. 'Down' places a 15 dB loss in the input.
SW2	30 dB volume control "capacitive" attenuator. 'Down' places a 30 dB loss in the input.
SW3:	'Up' position for operation in the LoLo (520-700) and HiLo (700-943 kHz) band. 'Down' position for operation in the LoHi (943-1270) and HiHi (1270-1710 kHz) band.
SW4:	Same as SW3.
SW5:	'Up' position for operation in the LoLo and LoHi band. 'Down' position for operation in the HiLo and HiHi band.
SW6:	'Up' position for normal crystal radio set operation, using a diode having an Is of about 100 nA. 'Down' position for increased selectivity, using a diode having an Is of about 15 nA in the #2 position, or for comparison testing of diodes.
SW7:	'Down' position for normal operation. 'Up' position to bypass the onboard audio transformers, if one wishes to use an external one.

The diode: This design is optimized for use with a diode having an n of 1.03 and a Saturation Current (Is) of about 82 nA, although this is not critical and other diodes can be used with good results. See Articles #0, #4 and #16 for info on n and saturation current (Is) of diodes, and how to measure them. If desired, and one has a favorite diode, its effective (Is) can be changed by applying a DC bias voltage, using perhaps, the 'Diode Bias Box' described in Article #9.

One suitable diode, the published parameters of which show an (Is) of 100 nA is a Schottky diode, the Agilent HBAT-5400. It is a surface-mount unit that was originally designed for transient suppression purposes. Measurements of many HBAT-5400 diodes seems to show that there are two varieties. One type measures approximately: n=1.03 and (Is)=80 nA. The other type seems to have an n of about 1.16 and an (Is) of

single or double-tuned Benodyne, it is suggested that the design be based on version "c" described in this Article, and not version "b", described in Article #22.

#26 Published: 02/10/2003; Revised: 10/23/2006

resistance, and the audio output resistance of the detector diode is about 2 times the RF source resistance driving it. The conditions for weak signal reception are different. Both the input RF and the output audio resistance of the diode detector are equal to the axis-crossing resistance of the diode. This resistance is Rd=0.026*n/Is (see Article #0 for a discussion on this). The compromise selected was to make the audio load resistance about 325k ohms. The design is optimized for a compromise diode having an Is of about 100 nA and an n of about 1.03. This diode has an axis-crossing resistance of 265k ohms. Most ITT FO-215 diodes have this characteristic. Many others can be used, with little effect on strong signal performance. The diode has more effect on weak than on strong signal performance. A ferrite stick inductor could do well for the coil in the antenna tuner (unit #1) if it has a high enough Q.

Bear in mind that the secondary coil should be constructed with the 250/62.5 uH contra wound arrangement and the 0.707 voltage taps. For this unit, the detector tuner, a 365 pF variable cap would be OK; the extra capacitance of a 485 pF capacitor is not necessary. The inductor for the antenna tuner should be constructed using the contra wound 250/62.5 uH arrangement, but no 0.707 voltage taps are needed.

It is possible to build a double-tuned Benodyne using primary and secondary coils of unequal values. I chose using equal values. A higher inductance for the primary might give some impedance matching problems at the low end of the LoLo band. A lower inductance for the secondary would require a lower RF loading resistance from the diode, thus requiring a diode having a higher Is. Look at Fig. 2 in Article #27 to see how weak signal sensitivity is reduced when using diodes having higher values of Is. For anyone interested in building a

about 150 nA. Both work well but the former works the best. This part, available in an SOT-23 package is easily connected into a circuit when soldered onto a "Surfboard" such as manufactured by Capital Advanced Technologies (http://www.capitaladvanced.com/), distributed by Alltronics, Digi-Key and others. Surfboard #6103 is suitable. The HBAT-5400 is also available in the tiny SO-323 package that can be soldered to a 330003 Surfboard.

The Agilent HSMS-2860 microwave diode (Specified Is=50 nA) is available as a single or triple with three independent diodes in the SO-323 and SO-363 packages, respectively. The Agilent number for the triple diode is HSMS-286L. I find it to be particularly good for DX in this crystal radio set. It is a convenient part since one can connect it using only one section (shorting the unused ones) or with two or three in parallel. This gives one a choice of nominal saturation currents of 50, 100 or 150 nA. Samples of this part I have tested measured about 35 nA per diode, not 50. I don't know the normal production variations. The only disadvantage of this diode, as far as I know, is its low reverse breakdown voltage which may cause distortion and low volume on very loud stations. It has the advantage, as do most Schottkys, of having much less excess reverse leakage current than do germanium diodes. This helps with volume and selectivity on very weak stations.

Infineon makes a BAT62 Schottky diode in several different quite small surface mount packages. The single BAT62 is physically the largest. It has a specified (Is) of about 100 nA and performs quite well. Be forewarned that the diode parasitic series resistance is a high 100 ohms in this diode. A resistance even this high should not have a noticeable effect on the performance of a crystal radio set.

Most germanium diodes have too high a saturation current for the best selectivity and should to be back-biased or cooled for optimum performance. See Article #17A for more info on this. Different type diodes may be connected to the terminals labeled Diode #1 and Diode #2, with either one selectable with SW6. When one diode is selected, the other is shorted. This feature makes it easy to compare the performance of a 'test' diode with one's 'favorite' diode. Another use is to place one's best DX diode in one position and one having a very low reverse leakage resistance at high reverse voltages in the other. This will maximize strong signal volume and minimize audio distortion.

A good choice for this crystal radio set is a diode having a relatively low saturation current such as 3 or 4 Agilent HSMS-2820 or HSMS-2860 diodes in parallel as Diode #1 for high selectivity and sensitivity on weak signals, and an Agilent HBAT-5400 or one of the lower saturation current germaniums as Diode #2 for low distortion and maximum volume on strong stations. Don't use two diodes in series if you want the best weak signal sensitivity in any crystal radio set. The result of using two identical diodes in series is the simulation of an equivalent single diode having the same (Is) but an n of twice that of either one. This reduces weak signal sensitivity.

The inductor for this single tuned crystal radio set is made up of the four closely coupled inductors L1, L2, L3 and L4. The inductance from point A to ground (Fig. 5) is 250 uH when SW3 is in the 'up' position (used for low band reception) and 62.5 uH in the 'down' position (used for high band reception). Better performance from a higher tank Q at the high frequency end of band B may be obtained by using the "contra-wound" coil winding technique described in Article #26. This

the crystal radio set in Article #22, is to enable an adjustable impedance transformation (vs. Frequency) between the antenna-ground system source resistance (assumed to be 25 ohms) and the top of the tank (point A). Since the primary and secondary inductors are of the same value and their loaded Q values are designed to be equal, the transformed antenna-ground resistance at the top of the primary tank should be made the same as that at the top of the secondary tank, including loading from the diode. These values are approximately (at band center): LoLo band:250k; HiLo band:500k; LoHi band:500k; HiHi band:1000k. Since the diode RF load is tapped at the 0.707 voltage point for bands HiLo and HiHi, the RF resistance driving the diode is in these cases is 250k and 500k respectively, the same as in the case of the LoLo and HiLo bands. If the diode was not tapped at the 0.707 voltage point for bands HiLo and HiHi, their band-center -3 dB bandwidths would be twice as wide as those of bands LoLo and LoHi. This equalizes all four band-center bandwidths to the desired value. If the tank inductance were the same for the four bands, the -3dB bandwidth at the center of the LoHi and HiHi bands would be four times as great as that at the center of the LoLo and HiLo bands. To correct this condition, the tank inductance in bands LoHi and HiHi is reduced from the 250 uH used in bands LoLo AND HiLo to 62.5 uH. This reduces the -3dB bandwidths at the center of bands LoHi and HiHi so they are the same as the bandwidths at the center of the LoLo and HiLo bands. There is some conflict in the design of a crystal radio set for best performance on strong signals (well above the "Linear-to-square-law crossover point), and one designed for weak signals (well below the "Linear-to-square-law crossover point), when one uses the same diode and audio transformation in both instances. This is because for strong signals, the RF input resistance of the diode detector is about half the audio load and band B. The steps I took to reduce these losses are: (1) Mounting C7 to the baseboard using strips of 0.10" thick, 0.5" wide and 1.5" long high-impact styrene as insulators and aluminum angle brackets screwed to the baseboard and (2), wiring these brackets to ground. This electrically isolates the capacitor formed from the lossy dielectric of the wooden baseboard from the rotor of C7. Ceramic stand-off insulators can be adapted, in place of the styrene strips for the job. Another way to mount C7 is to make a mounting plate from a sheet of low loss dielectric material, somewhat larger than C7's footprint, and screw C7 on top of it. Other holes made in the plate can then be used, along with small brackets or standoffs to mount the assembly onto the baseboard. Don't forget to wire the metal mounting pieces screwed to the wood baseboard to ground. These same considerations apply to any metal coil mounting bracket, close to a hot end of the coil, used to mount the coil form to the baseboard. The bracket should be grounded to short out its capacitance through the wood baseboard to ground. The contra wound coil configuration used in this crystal radio set is very helpful here since both outside ends of the coil, in band B are at ground potential.

12. Appendix: Design approach for double and single tuned Benodyne versions, as posted to the Discussion Group Rap 'n Tap (edited).

Ben H. Tongue Posted - 25 February 2005 15:53; The Benodyne approach to a double-tuned crystal radio set is shown in Article #23. The basic circuit is that of a conventional double-tuned circuit operating with equal loaded-Q values for primary and secondary. I decided to use equal values of inductance for the primary and secondary coils for convenience. The reason to use the two varicaps, C7 and C8 (in the primary circuit in unit #1), connected as they were in

minimizes distributed coil capacitance in band B as opposed to the winding connections used here that minimize coil distributed capacity in band A.

Audio impedance transformation from the audio output resistance of the diode detector to 'series connected' 1.2k ohm sound-powered phones is provided by the audio transformers. If one wishes to use 300 ohm sound-powered phones with two 600 ohm elements connected in parallel instead of series, a very good low loss transformer choice is the 100k-100 ohm transformer from Fair Radio Sales, #T3/AM20. The configuration of two Calrad transformers shown on line 2 of the Calrad chart in Article #5 is also a good choice. C11, along with the shunt inductance of the transformer and the inductance of the sound powered phones form a high-pass filter, hopefully flat down to to 300 Hz. R3 is used to adjust the DC resistance of the diode load to the AC impedance of the transformed effective AC headphone impedance. C10 is an audio bypass.

The two variable capacitors C7 and C8 interact substantially when tuning a station. C7 mainly controls the selectivity and C8 mainly controls the resonant frequency. If the antenna-ground system being used has a resistance larger than 25 ohms, C7 will have to be set to a smaller capacitance in order to maintain the proper resonant resistance at point A in Fig. 5. If the capacitance of the antenna-ground system is greater than 200 pF, C7 will also have to be set to a lower value than if it were 200 pF.

The "capacitive" attenuators controlled by SW1 and SW2, used for volume and selectivity control, are designed so as to cause minimal tank circuit detuning when the equivalent circuit of the antenna-ground system used has the same values as the old IRE simplified Dummy Antenna recommended for testing Broadcast Band radio receivers. It consisted of a series combination of a 200 pF cap, 20 uH inductor and a 25 ohm resistance. The geometric mean of the sum of the reactances of the capacitor and inductor at 520 and 1710 kHz is (-605) ohms. This is the reactance of a 279 pF capacitor (characteristic capacitance of the "capacitive" attenuator) at 943 kHz, the geometric mean of the BC band of 520-1710 kHz. The "capacitive" attenuators were designed for the specified attenuation values (15 and 30 dB) utilizing the 500 ohm resistive pi attenuator component values table shown in the book "Reference Data for Radio Engineers". The resistor values for 15 and 30 dB "capacitive" attenuators were normalized to 605 ohms, then the "capacitive" attenuator capacitor values were calculated to have a reactance, at 943 kHz, equal to the value of the corresponding "capacitive" attenuator shunt or series resistance. Since the "capacitive" attenuators, when switched into the circuit, isolate the antenna-ground system resistance from the tank circuit, selectivity is increased. This is a convenient feature, since less retuning is required than if selectivity is increased by reducing C7 and increasing C8. If the series capacitance of the equivalent circuit of one's own antenna-ground system is 200 pF, at 943 kHz, practically no retuning is required.

If the equivalent L and C of one's own antenna-ground system differ substantially from those of the simplified IRE dummy antenna used here, one can normalize the values of the capacitors used in the "capacitive" attenuators to match ones's own antenna-ground system. A method for measuring the parameters of one's own antenna-ground system is shown in Article #20.

cap. At lower frequencies, the dielectric material of the coil form becomes less important since its contribution to the distributed capacity is swamped out by the larger capacitance needed from the tuning capacitor in order to tune to the lower frequencies.

11. Important information re: maximizing unloaded tank Q, especially at the high end of the band.

Every effort should be made to achieve as high an unloaded tank Q as possible in order to minimize RF loss at the desired -3 dB bandwidth (selectivity), and especially when using narrower bandwidths. Somewhat greater insertion power loss and/or broader selectivity may result if components having a greater dielectric loss than those specified are used. Sensitive areas for loss are:

1. Q of the coil. See Table 3 for the Q values realized in the tank circuit.
2. Stator insulation material used in the variable caps C7, C8. Very important! Ceramic is best.
3. Skin-effect resistive loss in the variable capacitor plates. Silver plated capacitor plates have the least loss, brass or cadmium plated plates cause more loss. Aluminum plates are in-between. Rotor contact resistance can be a problem.
4. The type of plastic used in slide switches SW1, 2, and 4.
5. Front panel material.
6. Coil form material. High impact styrene has less dielectric loss than PVC. Styrene forms are available from Genova Products: http://genovaproducts.com/factory.htm . The forms are listed as drain couplers in their "400" series of products.
7. Capacitive coupling from any hot RF point, through the wood base to ground must be minimized because it tends to be lossy and will reduce performance at the high end of band A

5 divisions. This closely represents a 3 dB reduction in signal. Record those frequencies. You may encounter some hum and noise pickup problems and will have to respond appropriately to eliminate them. It is usually beneficial to conduct experiments of this type over a spaced, grounded sheet of aluminum placed on top of the workbench.

5. Calculate approximate unloaded tank Q. Qa=fo/(fh-fl). Calculate the actual Q by dividing Qa by 1.02 to reflect the fact that 5/7 does not exactly equal SQRT (0.5).

6. Try reducing the loop magnetic and probe capacitive coupling, and repeat the measurement and calculation. If the Q comes out about the same, that shows that the 50 output resistance of the generator and the scope probe loading do not significantly load the tank.

7. Note: When measuring the Q of an inductor with a Q meter it is common practice to lump all of the losses into the inductor. This includes magnetic losses in the inductor as well as dissipative losses in its distributed capacitance. We generally try to get a grip on tank Q values by measuring the inductor with a Q meter, when one is available. We assume that all the loss that affects the measured Q is magnetic loss. Not so, there is also loss in the dielectric of the distributed capacitance of the inductor. Actually, we are measuring an inductor having a specific Q (at a specific frequency), in parallel with the distributed capacity of the coil. We usually assume that the Q of this distributed capacity is infinite, but it isn't. The dielectric of the coil form material makes up much of the dielectric of the coil's distributed capacity and is the controlling factor in causing different coil Q readings when using coil forms made up of various different materials. This distributed capacity is in parallel with the tuning capacitor and can have an important effect on overall tank Q at the high end of the band because there, it is paralleled with the small, hopefully high Q, capacitance contribution from the variable

4. 'Loop Effect' of the tank inductor, and how it can be used to tame
local 'Blowtorch' stations when searching for DX.

One can use local signal pickup by the tank (loop effect) to reduce the effect of interference from strong stations by rotating the crystal radio set about a vertical axis. The correct angle will generally reduce it.

5. How to improve selectivity with a relatively small loss in sensitivity.

*

Selectivity can always be increased by reducing the value of C7 and re-tuning C8. If neither "capacitive" attenuator is in-circuit, switching one into the circuit will increase selectivity (and reduce volume).

*

Selectivity can be increased by changing to a diode having a lower Is than the HBAT-5400, such as the Agilent 5082-2835 or HSMS-2820. A DC bias, applied to the 'Diode Bias' terminals can 'fine-tune' performance. The diode 'Bias Box' described in Article #9 is useful here. One can choose less audio distortion and less selectivity by biasing the diode in a more forward direction, or better selectivity, at the cost of more audio distortion by biasing the diode toward its reverse direction.

*

Selectivity in the LoLo band (520-700 kHz) can be increased somewhat from the performance resulting from using the settings shown in Table 1 by switching SW5 to the 'down' position, and even more by, in addition, switching SW4 to the 'down' position.

*

Selectivity in the HiLo band (700-943 kHz) can be increased from the performance resulting from using the settings shown in Table 1 by switching SW4 to the 'down' position.

*

Selectivity in the LoHi band (943-1270 kHz) can be increased somewhat from the performance resulting from using the settings shown in Table 1 by switching SW5 to the 'down' position.

*

The only way to increase selectivity in the HiHi band is to use a diode having a low Is, reducing C7, or switching in a "capacitive" attenuator such as SW1 or SW2. See Fig. 5.

*

A large increase in selectivity can be attained by going to a double tuned circuit. See Article #23.

Note: When altering selectivity by changing switch positions, always re-balance the relative settings of C7 and C8.

6. Just how loud is a station that delivers the amount of power necessary to operate the Diode Detector at its 'Crossover Point' between Linear and Square-Law Operation?

Many Articles in this series have talked about the 'Linear-to-Square-Law Crossover Point' (LSLCP). Please bear in mind that the LSLCP point is a point 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. Two things can be said about a detector when it is fed a signal that operates it at its LSLCP. (1) A moderate increase of signal power will move the detector into its region of substantially linear operation. (2) A similar moderate decrease of input power will move it closer to its region of substantially square law

*A diode detector is operating at its LSLCP when its DC load resistance is R=n*0.0257/Is and the detected DC across it is 0.051 volts.

This CW method of measuring loss is much easier than the more complicated general method using AM modulation, as shown in Article #11.

10. A method for measuring the unloaded Q of an L/C resonator

1. Connect the 50-ohm output of a precision frequency calibrated RF generator (I used an Agilent digitally synthesized unit.) to a radiating test loop by means of, say, a 5 foot long coax cable. The loop can be made from 15 turns of solid #22 ga. vinyl insulated wire, bunched up into a ¼" diameter cross section bundle, wound on a 2" diameter vitamin pill bottle. The coil is held together by several twist-ties.
2. Make sure that all resistive loads are disconnected from the tank. Remove all potentially lossy non-metallic and metallic (especially ferrous) material from the vicinity of the coil. Capacitively couple the probe of a 5 MHz (or greater) scope to the hot end of the L/C tank and set the probe to its 1:1, not its 10:1 setting. This coupling must be very weak. This can be done by clipping the scope probe onto the insulation of a wire connected to the hot end of the coil (or a tap) or placing the probe very close to the hot end.
3. Place the 2" loop on-axis with the coil, about 6" from its cold (grounded) end. Tune the generator to fo MHz and adjust the generator output, scope sensitivity and L/C tuning to obtain, say, a 7 division pattern from fo on the scope. Note the frequency.
4. Detune the generator below and then above fo to frequencies (fl and fh) at which the scope vertical deflection is

The data in Table 5 show the insertion power loss when the crystal radio set is driven by a CW RF signal, with the diode feeding a resistive load of 355k. R3 is used for the resistive load and is set to 355k (be sure to disconnect any diode connected to the terminals when setting R3 for 355k). SW6 is set to the 'down' position. For greatest measurement accuracy, one should short out the series connected primaries of T1 and T2. The signal generator used in the measurements is adapted to have a source impedance equal to that of the standard IRE simplified dummy load (see Article #11). The expected diode detector power loss, at the output power level used, is about 5 dB. The remainder of the insertion loss shown in Table 5 is caused by losses in the inductive and capacitive parts of the tank. Weaker signals will result in a higher detector power loss, stronger signals, a lower loss. Some more info on detector power loss and LSLCP is given in Fig. 2 and its succeeding paragraph in Article #15a.

The signal level used was chosen to operate an HBAT5400 diode having an Is of 106 nA and an n of 1.03 near its LSLCP. The input RF voltage was set to cause a DC output voltage of 0.075 volts* across the R3=355k ohms (measured at the 'Diode DC V.' terminals). This gives an output power of -78 dBW. The signal generator output was varied, depending on the insertion power loss of the passive components as measurements were made at the different frequencies. In practice, one should add about 1.0-2.0 dB to the insertion power loss shown in Table #5 to allow for a typical audio transformer loss. In actual practice, of course, one uses an audio load (headphones), fed through the audio transformer, instead of a resistor for the diode load.

operation, where a 1 dB decrease of input power results with a 2 dB decrease of output power. For more info on the LSLCP, see Article #15A.

The crystal radio set described in this Article is operating at its LSLCP if the rectified DC voltage at the 'Diode Bias' terminals is 53 mV, a diode having a Saturation Current (Is) of 82 nA and an ideality factor (n) of 1.03 is used (such as a selected Agilent HBAT-5400), and if R3 is set to 325k ohms. At this point the diode rectified current equals two times its Saturation Current. The volume obtained is usually a low to medium, easy-to-listen-to level.

7. 'Short Wave ghost Signal', 'background hash' and spurious FM reception.

All single tuned crystal radio sets may be, in fact, considered double tuned (except single tuned loop receivers). The second response peak arises from resonance between the equivalent inductance of the antenna-ground system and the impedance it sees, in this case, the series combination of capacitors C7 and C8. This peak usually appears at a frequency above the broadcast band and gives rise to the possibility of strong so-called 'short wave ghost' signal interference when a short wave station has a frequency near the peak . The response at this ghost frequency can be made somewhat weaker and moved to a higher frequency if the antenna-ground system inductance is reduced. One can use multiple spaced conductors for the ground lead to reduce its inductance. I use a length of TV 300 ohm twin lead, the two wires connected in parallel for this purpose. Large gauge antenna wire, or spaced, paralleled multiple strands helps to reduce the antenna inductance (flat top antenna). If the down-lead is long compared to the ground lead, use multiple, paralleled, spaced conductors to reduce its inductance (similar to using a 'cage' conductor).

Another possible cause of 'ghost' signal reception resides in the fact that the response of the so-called single tuned circuit does not continuously drop above resonance as frequency rises, but only drops to a relatively flat valley before rising again to the second peak. The frequency response above the main (lower) peak would drop monotonically (true single tuned operation) if the second peak did not exist. The relatively flat response valley that exists between the two peaks, provides the possibility (probably likelihood) of interference 'hash' if several strong stations are on the air at frequencies in the valley range. It also is the cause of a strong local station, above the frequency of a desired station "riding through" and appearing relatively constant even if the tuning dial is moved. The response should drop at a 12 dB per octave rate above the second peak. A useful side effect of the response behavior of this type of circuit is that the response below the main resonance drops off at an extra fast rate of 12 dB per octave rate instead of an expected 6 dB.

The most effective way to substantially eliminate 'short wave ghost' and hash reception is to go to a double-tuned circuit configuration or to use traps.

Spurious FM reception caused by so-called FM 'slope' detection can occur from close by local FM stations if a spurious FM resonance appears somewhere in the circuitry of a crystal radio set. If ground wiring is not done properly in the crystal radio set, spurious signals can get into the detector. The thing to do here is to run all the RF and audio grounds to one point as shown in Fig.5. Sometimes a small disc bypass capacitor, 22 pF or so, placed across the diode will help.

Fig. 11 shows the simulated frequency response at the center of sub-band 4, from the antenna-ground system source to the RF input of the diode. The red graph and figures in the left panel show an insertion power loss of 4.1 dB with a -3 dB bandwidth of 6 kHz, along with the spurious response peak at 6.9 MHz, caused by the antenna-ground system inductance. The insertion loss at the spurious peak is 20 dB. The loss in the valley is 47 dB. The right graph and figures show the Input Return Loss (impedance match) at resonance to be -8.5 dB. The output return loss is the same.

Fig. 10 and Fig.11 are actually simulations of the RF frequency response of version 'b' as described in Article #22. The response curves of version 'c', described in this article should be the same, with the exception of halving less loss at the peak response points. This is because of the higher tank Q in version 'c'.

Table 4 - Measured unloaded tank Q values. Antenna and diode disconnected,

Table 4 - Measured unloaded tank Q values. Antenna and diode disconnected, SW1 and SW2 set to -15 and -30 dB and C7 set to 50 on the dial

Band-->	Sub-band 1	Sub-band 2	Sub-band 3	Sub-band 4
Frequency in kHz.-->	520	943	943	1710
Measured unloaded tank circuit Q (includes loss in the tuning caps, switches and all other misc. loss)-->	1020	1000	1240	940

Table 5 - Measured RF bandwidth and power loss @ resonance, at approx. the LSLCP. Diode parameters: Is=106 nA, n=1.03. Output power=15.8 nW=-78 dBW

Center frequency in kHz	Insertion loss in dB, normal selectivity	-3 dB bandwidth in kHz, normal selectivity	Insertion loss in dB, sharp selectivity	-3 dB bandwidth in kHz, sharp selectivity
603	6.5	6.3	9.0	1.9
813	7.3	5.0	10.0	2.2
1094	7.5	4.7	12.0	2.2
1474	8.8	5.8	13.5	2.9

Table 5 - Measured RF bandwidth and power loss @ resonance, at approx. the

9. Measurements.

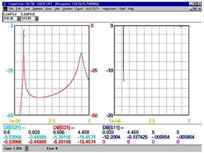


Fig. 10 - RF frequency response from antenna to diode input, center of sub-band 1, normal selectivity, using the simplified IRE dummy antenna.

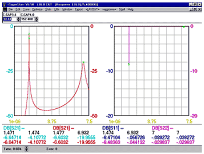


Fig. 11 - RF frequency response from antenna to diode input, center of sub-band 4, normal selectivity, using the simplified IRE dummy antenna.

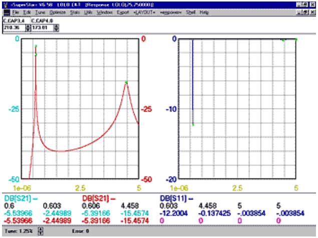
Response graph of LoLo band
Fig. 10 - RF frequency response from antenna to diode input, center of sub-band 1, normal selectivity, using
the simplified IRE dummy antenna.

Respomse graph of HiHi band
Fig. 11 - RF frequency response from antenna to diode input, center of sub-band 4, normal selectivity, using
the simplified IRE dummy antenna.

Fig. 10 shows the simulated frequency response at the center of sub-band 1, from the antenna source to the RF input of the diode. The red graph and figures in the left panel show an insertion power loss of 2.4 dB with a -3 dB bandwidth of 6 kHz, along with the spurious response peak at 4.4 MHz, caused by the antenna-ground system inductance. The insertion loss at the spurious peak is 15 dB. The loss in the valley is 40 dB. The right graph and figures show the Input Return Loss (impedance match) at resonance to be -12.2 dB. The output return loss (not shown) is the same.

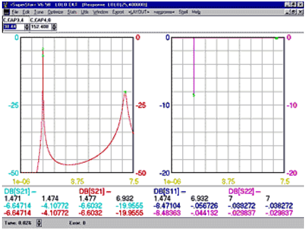
Another way to try to reduce FM interference is to put a wound ferrite bead 'choke' in series with the antenna and/or ground leads. In order not to affect normal BC band reception, the resultant ferrite inductor should have a reasonable Q and a low inductance in the BC band. It should also exhibit a high series resistance at FM frequencies. Suitable wound ferrite chokes (bead on a lead) are made by the Fair-Rite Corp. as well as others. Two types available from Mouser are #623-29441666671 and #623-2961666671. This suggestion may also help reduce "short wave ghost" signal reception.

8. Some simulated and actual measurements on the crystal radio set.



Response graph of LoLo band

Fig. 7 - RF frequency response from antenna to diode input, center of LoLo Band, using the simplified dummy antenna.



Respomse graph of HiHi band

Fig .8 - RF frequency response from antenna to diode input, center of HiHi Band, using the simplified dummy antenna.

Fig. 7 shows the simulated frequency response at the center of the LoLo band, from the antenna source to the RF input of the diode. The red graph and figures in the left panel show an insertion power loss of 2.4 dB with a -3 dB bandwidth of 6 kHz, along with the spurious response peak at 4.4 MHz, caused by the antenna-ground system inductance. The insertion loss at the spurious peak is 15 dB. The loss in the valley is 40 dB. The right graph and figures show the Input Return Loss (impedance match) at resonance to be -12.2 dB. The output return loss (not shown) is the same.

Fig. 8 shows the simulated frequency response at the center of the HiHi band, from the antenna source to the RF input of the diode. The red graph and figures in the left panel show an insertion power loss of 4.1 dB with a -3 dB bandwidth of 6 kHz, along with the spurious response peak at 6.9 MHz, caused by the antenna-ground system inductance. The

appearing relatively constant even if the tuning dial is moved. The response should drop at a 12 dB per octave rate above the second peak. A useful side effect of the response behavior of this type of circuit is that the response below the main resonance drops off at an extra fast rate of 12 dB per octave rate instead of an expected 6 dB.

The most effective way to substantially eliminate 'short wave ghost' and hash reception is to go to a double-tuned circuit configuration.

Spurious FM reception caused by so-called FM 'slope' detection can occur from close by local FM stations if a spurious FM resonance appears somewhere in the circuitry of a crystal radio set. If ground wiring is not done properly in the crystal radio set, spurious signals can get into the detector. The thing to do here is to run all the RF and audio grounds to one point as shown in Fig.5. Sometimes a small disc bypass capacitor, 22 pF or so, placed across the diode will help.

Another way to try to reduce FM interference is to put a wound ferrite bead 'choke' in series with the antenna and/or ground leads, if the FM interference is coming in on those leads. In order not to affect normal BC band reception, the resultant ferrite inductor should have a reasonable Q and a low inductance in the BC band. It should also exhibit a high series resistance at FM frequencies. Suitable wound ferrite chokes (bead on a lead) are made by the Fair-Rite Corp. as well as others. Two types available from Mouser are #623-29441666671 and #623-2961666671. This suggestion may also help reduce "short wave ghost" signal reception in some cases.

Note: See "Wiring the crystal radio set:", in Part 3, above.

All single tuned crystal radio sets may be, in fact, considered double tuned (except single tuned loop receivers). The second response peak arises from resonance between the equivalent inductance of the antenna-ground system and the impedance it sees, in this case, the series combination of capacitors C7 and C8. This peak usually appears at a frequency above the broadcast band and gives rise to the possibility of strong so-called 'short wave ghost' signal interference when a short wave station has a frequency near the peak . The response at this "ghost" frequency can be made somewhat weaker and moved to a higher frequency if the antenna-ground system inductance is reduced. One can use multiple spaced conductors for the ground lead to reduce its inductance. I use a length of TV 300 ohm twin lead, the two wires connected in parallel for this purpose. Large gauge antenna wire, or spaced, paralleled multiple strands helps to reduce the antenna-ground system inductance (flat top antenna). If the down-lead is long compared to the ground lead, use multiple, paralleled, spaced conductors to reduce its inductance (similar to using a 'cage' conductor).

Another possible cause of 'ghost' signal reception resides in the fact that the response of the so-called single tuned circuit does not continuously drop above resonance as frequency rises, but only drops to a relatively flat valley before rising again to the second "ghost" peak. The frequency response above the main (lower) peak would drop monotonically (true single tuned operation) if the second peak did not exist. The relatively flat response valley that exists between the two peaks, provides the possibility (probably likelihood) of interference 'hash' if several strong SW stations picked up at frequencies in the valley range. This also is the cause of a strong local station, above the frequency of a desired station "riding through" and

insertion loss at the spurious peak is 20 dB. The loss in the valley is 47 dB. The right graph and figures show the Input Return Loss (impedance match) at resonance to be -8.5 dB. The output return loss is the same.

Table 2 - Expected and Measured Tank Q values
(antenna and diode disconnected)

Band	LL	HL	LH	HH
Frequency in kHz	603	813	1094	1474
Expected coil Q, according to Medford	497	577	669	777
Measured, unloaded tank circuit Q (includes loss in the tuning caps and all other misc. loss)	431	463	555	620

9. A method for measuring the unloaded Q of an L/C resonator.

1.

Connect the 50-ohm output of a precision frequency calibrated RF generator (I used an Agilent digitally synthesized unit.) to a radiating test loop by means of, say, a 5 foot long coax cable. The loop can be made from 15 turns of solid #22 ga. vinyl insulated wire, bunched up into a ¼ " diameter cross section bundle, wound on a 2" diameter vitamin pill bottle. The coil is held together by several twist-ties.

2.

Make sure that all resistive loads are disconnected from the tank. Remove all metallic (especially ferrous) material from the vicinity of the coil. Capacitively couple the probe of a 5 MHz scope to the hot end of the L/C tank. This coupling must be very weak. This can be done by clipping the scope probe onto the insulation of a wire connected to the hot end of the coil (or a tap) or placing the probe very close to the hot end.

3.

Place the 2" loop on-axis with the coil, about 6" from its cold (grounded) end. Tune the generator to say, fo MHz and adjust the generator output, scope sensitivity and L/C tuning to obtain 7 division pattern from fo on the scope. Note the frequency.

4.

Detune the generator below and then above fo to frequencies (fl and fh) at which the scope vertical deflection is 5 divisions. This represents an approximate 3 dB reduction in signal. Record those frequencies. You may encounter some hum and noise pickup problems and will have to respond appropriately to eliminate them. It is usually beneficial to conduct experiments of this type over a spaced, grounded sheet of aluminum placed on top of the workbench.

5.

Calculate approximate unloaded tank Q. Qa=fo/(fh-fl). Calculate the actual Q by dividing Qa by 1.02 to reflect the fact that 5/7 does not exactly equal SQRT (0.5).

6.

Try reducing the loop and capacitive coupling, and repeat the measurement and calculation. If the Q comes out about the same, that shows that the 50 output resistance of the generator and the scope loading do not significantly load the tank.

7.

Note: When measuring the Q of an inductor with a Q meter it is common practice to lump all of the losses into the inductor. This includes magnetic losses in the inductor as well as dissipative losses in its distributed capacitance. We generally try to get a grip on tank Q values by measuring the inductor with a Q meter, when one is available. We assume that all the loss that affects the measured Q is magnetic loss. Not so, there is also loss in the dielectric of the distributed capacitance of the inductor. Actually, we are measuring an inductor having a specific Q (at a specific frequency), in parallel with the distributed capacity of the coil. We usually assume that the Q of this distributed capacity is infinite, but it

One can use local signal pickup by the tank (loop effect) to reduce the effect of interference from strong stations by rotating the crystal radio set about a vertical axis. The correct angle will generally reduce it.

7. Just how loud is a station that delivers the amount of power necessary to operate the
Diode Detector at its 'Crossover Point between Linear and Square Law Operation'?

Many Articles in this series have talked about the 'Linear to Square-Law Crossover' (LSLCP). Please bear in mind that the LSLCP is a point on a graph of DC output 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. Two things can be said about a detector when it is fed a signal that operates it at its LSLCP. (1) A moderate increase of signal power will move the detector into its region of substantially linear operation. (2) A similar moderate decrease of input power will move it closer to its region of substantially square law operation where a 1 dB decrease of input power results with a 2 dB decrease of output power.

The crystal radio set described in this Article is operating at its LSLCP if the rectified DC voltage at the 'Diode Bias' terminals is about 51 mV. This assumes a diode having an Is of 106 nA and an ideality factor of 1.03 (such as a selected Agilent HBAT5400 or an Infineon BAT62) is used, with R3 is set to 355k ohms. The audio volume obtained is usually a low to medium, easy-to-listen-to level when using sound power phones.

8. 'Short Wave ghost Signal', 'background hash' and spurious FM reception

to the same dial number and SW1 and SW2 are set to 0 and 30 dB, respectively.

Dial settings-->	0	10	20	30	40	50	60	70	80	90	100
Sub-band 1 switch setup, normal selectivity	386	423	**482**	**560**	**650**	**776**	867	994	1176	1394	1606
Sub-band 2 switch setup, normal selectivity	387	424	484	562	**658**	**765**	**881**	**1015**	1171	1389	1581
Sub-band 3 switch setup, normal selectivity	765	835	**951**	**1096**	**1268**	1466	1681	1921	2192	2565	2881
Sub-band 4 switch setup, normal selectivity	773	839	950	1104	**1278**	**1485**	**1708**	1953	2241	2650	2987

5. How to improve selectivity with a relatively small loss in sensitivity, in addition
to using the "Sharp Selectivity" switch positions on SW 4.

*

Selectivity can always be increased by reducing the value of C7 and retuning C8. If neither "capacitive" attenuator is in-circuit, switching one into the circuit will increase selectivity.

* Selectivity can be increased by changing to a diode having a lower Is than the HBAT5400, such as the Agilent 5082-2835 or HSMS-2820. A DC bias, applied to the 'Diode Bias' terminals can 'fine-tune' performance. The diode 'Bias Box' described in Article #9 is useful here. One can choose less audio distortion and less selectivity by biasing the diode in a more forward direction, or better selectivity, at the cost of more audio distortion by biasing the diode toward its reverse direction.

* Experimentation using a position on SW4 that taps the diode further down on the tank than specified in Table 2.

6. 'Loop Effect' of the tank inductor, and how it can be used to tame
local 'Blowtorch' stations when searching for DX.

isn't. The dielectric of the coil form material makes up much of the dielectric of the coil's distributed capacity and is the controlling factor in causing different coil Q readings when using coil forms made up of various different materials. This distributed capacity is in parallel with the tuning capacitor and can have an important effect on overall tank Q at the high end of the band because it is paralleled with the small, hopefully high Q, capacitance contribution from the variable cap. At lower frequencies, the dielectric material of the coil form becomes less important since its contribution to the distributed capacity is swamped out by the larger capacitance needed from the tuning capacitor in order to tune to the lower frequencies.

10. Important information re: unloaded tank Q.

Every effort should be made to achieve as high an unloaded tank Q as possible, in order to minimize RF loss at the desired -3 dB bandwidth (selectivity), and especially when using narrower bandwidths. Somewhat greater insertion power loss and/or broader selectivity may result if components having a greater dielectric loss than those specified are used. Sensitive areas for loss are:

1. Q of the coil. See Table 2 for the Q values realized in the tank circuit.
2. Stator insulation in the variable caps C7, C8.
3. Skin-effect resistive loss in the variable capacitor plates. Silver plated capacitor plates have the least loss, brass or cadmium plated plates cause more loss. Aluminum plates are in-between. Rotor contact resistance can be a problem.
4. Contact support plastic used in slide switches SW3, 4, 5 and 6.
5. Front panel material.

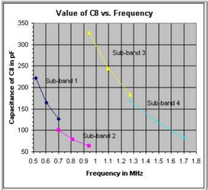
6. Coil form material. Styrene has 1/10 the dielectric loss of PVC. High impact styrene forms are available from Genova Products at their retail store: http://genovaproducts.com/factory.htm . These forms are listed as drain couplers.

7. Capacitive coupling from any hot RF point, through the wood base to ground must be minimized because it tends to be lossy and will reduce performance at the high end of band A and band B. The steps I took to reduce these losses are: (1) Mounting C7 to the baseboard using strips of 0.10" thick, 0.5" wide and 1.5" long high-impact styrene as insulators and aluminum angle brackets screwed to the baseboard and (2) and connecting these brackets to ground. This isolates the lossy dielectric of the baseboard from the hot end of C7. See Fig. 1. Ceramic stand-off insulators can be adapted, in place of the styrene strips for the job. Another way to mount C7 is to make a mounting plate from a sheet of low loss dielectric material, somewhat larger than C7's footprint, and screw C7 on top of it. Holes made in the plate can then be used, along with small brackets or standoffs to mount the assembly onto the baseboard. Don't forget to wire the metal mounting pieces to ground. These same considerations apply to any metal coil mounting bracket, close to a hot end of the coil, used to mount the coil form to the baseboard. The bracket should be grounded.

8. The physical size of the coil is important. A large size coil was chosen to enable a high Q. Medhurst's work enables one to calculate the Q of a solenoid wound with solid copper wire, provided that: 0.4<do/t<0.8. do=diameter of the wire, t=center-to-center spacing of the turns. If this relation is followed, for a given physical volume, the maximum Q will occur when D=L, where D=diameter of the coil and L=length of the coil. The Q is then proportional to D(=L). Much care is

It is assumed that the tank inductor has the required 250 and 62.5 uH inductance values and is contra wound as described. Of course different inductance values can be used to make a good crystal radio set, but the graphs in Figs. 8 and 9 would have to be changed. It is also assumed, when using the figs. 8 or 9, that the impedance of the antenna-ground system being used is equal to that of the Standard Dummy Antenna.

Graph of capacitance of C8 vs. frequency Graph of dial setting of C8 vs. frequency



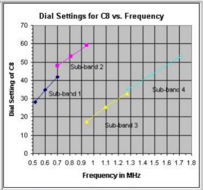


Fig. 8 - Designed capacitance of C8 vs. Frequency

Fig. 9 - Dial setting of C8 when using the specific variable capacitor specified in the Parts List.

Fig. 8 - Designed capacitance of C8 vs. Frequency

Fig. 9 - Dial setting of C8 when using the specific variable capacitor specified in the Parts List.

Another method of tuning is to estimate from Table 3 the dial settings for C7 and C8 required to tune to the desired station. C7 and C8 can then be tuned together higher or lower to actually tune in the station. If more selectivity is desired, reduce the capacitance of C7 and retune C8. If the volume is too low, try increasing C7 and decreasing C8.

Table 3 - Tuned frequency in kHz as a function of dial settings, if C7 and C8 are set

Table 2 - Switch Functions for Version C

SW1	15 dB volume control "capacitive" attenuator. 'Down' places about 15 dB loss in the input.
SW2	30 dB volume control "capacitive" attenuator. "Down' places about 30 dB loss in the input.
SW3	520-943 kHz - Band A (sub-bands #1 and #2) - position 1 943-1710 kHz - Band B (sub-bands #3 and #4) - position 2
SW4	520-700 kHz - Band A, sub-band 1. Normal selectivity: position 2, Sharp selectivity: position 3 700-943 kHz - Band A, sub-band 2. Normal selectivity: position 5, Sharp selectivity: position 4 943-1270 kHz - Band B, sub-band 3. Normal selectivity: position 2, Sharp selectivity: position 1 1270-1710 kHz - Band B, sub-band 4. Normal selectivity: position 4, Sharp selectivity: position 5
SW5	Used to select diode 1 or diode 2.
SW6	'Down' position for normal operation using 1.2k ohm phones. 'Up' position to bypass the onboard audio transformers if one wishes to use an external transformer to match an impedance other than 1.2k ohms.

4. Tuning the Crystal Radio Set to to a specific frequency.

C8 is considered the primary tuning control. C7 is used, in conjunction with the capacity of the antenna-ground system to adjust selectivity to the designed value. It also has considerable interaction with the tuning frequency. There are two methods for tuning in a station of a known frequency that will result in the selectivity being fairly close to specification, as shown in table 5. One requires a knowledge of the capacitance of C8 vs. its dial setting. The second requires fitting C8 with a knob having a linear calibration of 0-100 over a 180 degree span and having a dial reading of zero at maximum capacity. To use this method C8 must be as specified in the parts list or an exact equivalent.

1. To tune to a specific frequency, read the necessary capacity for C8 from Fig. 8 and set C8 to that value. Adjust C7 to tune in the station.

-or-

2. To tune to a specific frequency, read the necessary dial setting for C8 from Fig. 9 and set C8 to that value. Adjust C7 to tune in the station.

required in measuring the Q of physically large high Q coils. The method I favor is given in Part 9, above.

11. Measurements.

Table 3-Tuned frequency in kHz as a function of dial settings, if C7 and C8 are set
to the same dial number and SW1 and SW2 are set to 0 and 30 dB, respectively.

Dial setting:	0	10	20	30	40	50	60	70	80	90	100
LoLo band (27-47)	385	420	**473**	**548**	**630**	**738**	850	977	1115	1308	1488
HiLo band (46-67)	387	421	477	553	**644**	**749**	**869**	**1002**	1159	1378	1584
LoHi band (22-44)	754	818	**916**	**1054**	**1211**	**1385**	1579	1780	1996	2269	2498
HiHi band (43-65)	758	821	926	1062	**1225**	**1405**	**1601**	**1817**	2052	2355	2611

In use, C7 and C8 are usually set to different values to achieve the design-bandwidth of 6-7 kHz. However, if they are set to the same values, a frequency calibration chart can be made for each band as shown above. The bold figures indicate the approximate position of each band when the crystal radio set is driven by the standard antenna-ground system. There is sufficient extra capacitance range available in C7 and C8 to handle antenna-ground systems that differ substantially in impedance from the standard dummy antenna used in the design.

Table 4 - Measured RF bandwidth and insertion power loss, at an audio
output power of -70 dBW, using the method described in Article #11

Dial setting of C7, C8	Center frequency in kHz	Insertion power loss in dB	-3 dB bandwidth in kHz
35, 35	603	3.8	6.1
59, 53	813	4.5	6.4
68, 34	1094	4.8	7.3
77, 43	1474	4.6	7.5

The data in Table 4 shows the insertion power loss in the crystal radio set when driven by an RF signal that is amplitude modulated at 50% by a 400 Hz sine wave. See Article #11 on how to measure the insertion power loss and bandwidth of a crystal radio set. The audio output power was set to -70 dBW for each reading. The available carrier input power supplied to the crystal radio set was about -60 dBW, with a total available sideband power of about -66 dBW. The audio output power is that delivered by the diode detector to the audio load and does not include losses in an audio transformer. One should add about 1.0-1.5 dB to the insertion power loss shown in Table #4 to allow for audio transformer loss. The reason the audio transformer loss does not show up in the measurements is that the audio transformers (T1 and T2) were not used. SW7 was placed in the UP position, providing a direct high impedance output from the diode detector. The Zero Loss Unilateral 'Ideal Transformer' Simulator described in Article #14 was used to provide a 320k to 1200 ohm impedance transformation, close to that provided by T1 and T2 in actual crystal radio set operation.

Note: The diode rectified DC voltage at the power levels used above is 0.51 volts. At this power level, a SPICE simulation of the detector shows a theoretical diode detector insertion power loss of 1.4 dB.

#22 Published: 02/05/2002; Revised:08/20/2006

The "capacitive" attenuators controlled by SW1 and SW2: Used for volume and selectivity control and are are designed so as to cause minimal tank circuit detuning when the equivalent circuit of the antenna-ground system has the same values as the old IRE simplified Dummy Antenna recommended for testing Broadcast Band radio receivers. It consists of a series combination of a 200 pF cap, 20 uH inductor and a 25 ohm resistance. The geometric mean of the sum of the reactances of the capacitor and inductor at 520 and 1710 kHz is -605 ohms. This is the reactance of a 279 pF capacitor (characteristic capacitance of the "capacitive" attenuator) at 943 kHz, the geometric mean of the BC band of 520-1710 kHz. The "capacitive" attenuators were designed for the specified attenuation values (15 and 30 dB) utilizing the 500 ohm resistive pi attenuator component values table shown in the book "Reference Data for Radio Engineers". The resistor values for 15 and 30 dB "capacitive" attenuators were normalized to 605 ohms, then the "capacitive" attenuator capacitor values were calculated to have a reactance, at 943 kHz, equal to the value of the corresponding "capacitive" attenuator shunt or series resistance. Since the "capacitive" attenuators, when switched into the circuit, isolate the antenna-ground system resistance from the tank circuit, selectivity is increased. If the series capacitance of the equivalent circuit of one's own antenna-ground system is 200 pF, at 943 kHz, practically no retuning is required.

If the equivalent L and C of one's own antenna-ground system differ substantially from those of the simplified IRE dummy antenna used in this design, one can normalize the values of the capacitors used in the "capacitive" attenuators to match one's own antenna-ground system. A method for measuring the parameters of an antenna-ground system is shown in Article #20.

Audio impedance transformation: from the audio output resistance of the diode detector to 'series connected' 1.2k ohm sound-powered phones is provided by the audio transformers. If one wishes to use 300 ohm sound-powered phones with two 600 ohm elements connected in parallel instead of series, a very good low loss transformer choice is the 100k-100 ohm transformer from Fair Radio Sales, #T3/AM20. The configuration of two Calrad transformers shown on line 2 of the Calrad chart in Article #5 is also a good choice. C11, along with the shunt inductance of the transformer and the inductance of the sound powered phones form a high-pass filter, flat (hopefully) down to to 300 Hz. R3 is used to adjust the DC resistance of the diode load to the AC impedance of the transformed effective AC headphone impedance to minimize audio distortion on very strong signals. C10 is an audio bypass.

The two variable capacitors C7 and C8: interact substantially when tuning a station. C7 mainly controls the selectivity and C7 and C8 together control the resonant frequency. Reducing the capacitance of C7 increases selectivity. If the antenna-ground system has a resistance larger than 25 ohms, C7 will have to be set to a smaller capacitance in order to maintain the proper resonant resistance at point A in Fig. 5. If the capacitance of the antenna-ground system is greater than 200 pF, C7 will also have to be set to a lower value than if it were 200 pF. If the maximum capacitance of the capacitor used for C7 used is not large enough to enable a large enough bandwidth at the low end of sub-band 1, provision can be made to switch a 330 pF NPO ceramic cap in parallel with it. This may be needed if the antenna-ground system has too low a capacitance (small antenna).

ARTICLE 23

How to Make a Very Efficient Double-Tuned, Four-Band, MW Crystal Radio Set using two Version 'b' Single-Tuned, Four-Band, MW Crystal Radios

Quick Summary: A double-tuned four band crystal radio set (DT4BCS) can be created by coupling together two Version 'b' single tuned crystal radio sets (VbST4BCS). See Article #22.

Selectivity and sensitivity are essentially constant over the entire AM broadcast band because of the use of 'Benodyne' constant bandwidth antenna coupling, resonant impedance control and band splitting. The coupling coefficient of the resultant double tuned circuit is easily adjustable from greater than critical coupling to approximately zero. Insertion power loss varies from about 4.5 to 6.5 dB at the center of each of the four bands (including the diode detector loss but excluding audio transformer loss). Selectivity is quite sharp: -3 dB bandwidth is 6-7 kHz and -20 dB bandwidth varies from 16 to 22 kHz. The average ratio of the -20 dB to the -3 dB bandwidth is 3.0. This is the same as the theoretical value shown on a graph in Terman's Radio Engineer's handbook, page 160, for two critically coupled circuits resonant at the same frequency and having a Q ratio of unity. Operation can be quickly changed to that of a single tuned crystal radio set accompanied by a sharp and deep tunable trap.

1. Design approach.

The tank inductor in a VbST4BCS is mounted so that its axis makes a 30 degree angle with the front panel. This orientation enables the magnetic coupling between two VbST4BCS tuned circuits to be easily varied from above critical coupling down to about zero. Incidentally, if one uses two identical coils, wound in the same direction and positioned as recommended in Article #22, the capacity coupling between the coils will be phased such as to partially oppose the magnetic coupling. To create a DT4BCS, two VbST4BCS are placed side-by-side next to each other with baseboards touching each other. The coupling between the two tuned circuits can be varied by sliding one VbST4BCS forward or back compared with the other.

If one is starting from scratch and is only interested in having a double-tuned crystal radio set, unnecessary parts may be removed from each of the VbSt4BCS units as follows:

1. Antenna tuner unit: Eliminate switches SW4 through SW7 and all other parts to the right of SW4 as shown in Fig. 5 of Article #22.
2. Detector unit: Eliminate C7 and all parts to the left of it as shown in Fig. 5 of Article #22.

If one is interested in having a single-tuned crystal radio set as well as a double tuned one, one full VbST4BCS may be used with either of the reduced parts count units #1 or #2 above.

the difference is usually hard to notice. See Article #17A for more info on this.

Different type diodes may be connected to the terminals labeled Diode #1 and Diode #2, with either one selectable with SW5. When one diode is selected, the other is shorted. This feature makes it easy to compare the performance of a 'test' diode with one's 'favorite' diode. Another use is to place one's best DX diode in one position and one having a very low reverse leakage current at high reverse voltages in the other. This will maximize volume and minimize audio distortion on strong signals.

A good choice for this crystal radio set is a diode having a relatively low saturation current such as 3 or 4 Agilent HSMS-2820 or HSMS-2860 diodes in parallel as Diode #1 for high selectivity and sensitivity on weak signals, and an Agilent HBAT5400 or one of the lower saturation current germaniums as Diode #2 for low distortion and maximum volume on very strong stations. Don't use two diodes in series if you want the best weak signal sensitivity in any crystal radio set. The result of using two identical diodes in series is the simulation of an equivalent single diode having the same (Is) but an n of twice that of either one. This reduces weak signal sensitivity.

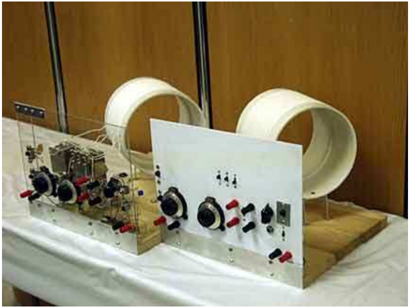
Different type diodes may be connected to the terminals labeled Diode #1 and Diode #2, with either one selectable with SW5. When one diode is selected, the other is shorted. This feature makes it easy to compare the performance of a 'test' diode with one's 'favorite' diode. Another use is to place one's best DX diode in one position and one having a very low reverse leakage resistance at high reverse voltages in the other. This will maximize strong signal volume and minimize audio distortion.

and an (Is) of about 150 nA Both work well but the former works best. This part, available in an SOT-23 package is easily connected into a circuit when soldered onto a "Surfboard" such as manufactured by Capital Advanced Technologies (http://www.capitaladvanced.com/), distributed by Alltronics, Digi-Key and others. Surfboard #6103 is suitable. The HBAT5400 is also available in the tiny SO-323 package that can be soldered to a 330003 Surfboard.

The Agilent HSMS-2860 microwave diode (Specified Is=50 nA) is available as a single or triple with three independent diodes in the SO-323 and SO-363 packages, respectively. The Agilent number for the triple diode is HSMS-286L. I find it to be particularly good for DX in this crystal radio set. It is a convenient part since one can connect it using only one section (shorting the unused ones) or with two or all three in parallel. This gives one a choice of nominal saturation currents of 50, 100 or 150 nA Samples of this part I have tested measured about 35 nA per diode, not 50. I don't know the normal production variations. The only disadvantage of this diode, as far as I know, is its low reverse breakdown voltage which may cause distortion and low volume on very loud stations. It has the advantage, as do most Schottkys, of having much less excess reverse leakage current than do germanium diodes. This helps with volume and selectivity on very weak stations.

Infineon makes a BAT62 Schottky diode in several different small surface mount packages. The single BAT62 is physically the largest and easiest to handle. It has a specified (Is) of about 100 nA and performs quite well.

Most germanium diodes have too high a saturation current for the best selectivity when receiving weak signals and should to be back-biased or cooled for optimum performance, although



Picture of double tuned set

Fig. 1 - A Double-tuned Four -Band Crystal Radio Set using a Version 'A' Single-tuned Crystal Radio Set as the antenna tuner/primary tuned circuit and a Version 'B' Single-tuned Crystal Radio Set for the secondary tuned circuit/detector function.

2. Operation of a DT4BCS when two VbST4BCS units are used.

Connect antenna and ground to unit #1. Position SW1 and SW2 at their 0 dB settings. Make sure no diode is connected to either the Diode #1 or Diode #2 terminals of unit #1. Connect phones to unit #2. Connect a detector diode to either of the two diode terminal pairs of unit #2 and switch it into the circuit using SW6.

In set #2, set SW1, SW2 to activate their attenuation, and C7 to minimum capacitance. Select one of the four bands for listening and set SW3, SW4 and SW5 in set #2 appropriately, as described in Table 1 in Article #22. Set SW3 in unit #1 to its up position when using bands LoLo and HiLo and its down position when and for using bands LoHi and HiHi.

Listening is usually done using critical coupling between the two tuned circuits. This occurs in the Lo band when the front panel of unit #2 is placed about 2 7/8" back from that of unit #1. The figure for the Hi band is about 3 3/4 ". Almost complete cancellation of coupling occurs when the front panel of unit #2 is pushed about 5 3/8" back from that of unit #1.

Tune in a station using C7 and C8 of unit #1 and C8 of unit #2. Greater selectivity is possible if needed. See Part 5 of Article #22 for more info on this. If interference is a problem, try reducing coupling below critical by moving set #2 further back. Very important: Optimum operation and best selectivity occurs when the loaded tank Q of unit #1 equals that of unit #2. A way to check for for this is to tune in a station and lightly place a finger on the stator (point A in Fig. 5 of Article #22) of C8 of unit #1, then on C8 of unit #2. If similar decreases of volume occur, the Qs are about equal. If the volume is reduced more by touching C8 of unit #1 than unit #2, increase C7 of unit #1 somewhat, restore tuning with C8 and try again. If the volume is reduced more touching C8 of unit #2 than unit #1, decrease C7 somewhat, restore tuning with C8 and try again for equal effects.

If a strong local station seems to 'bleed through' the tuned circuits, that may be because the inductor of unit #2 is acting as a loop antenna and picking it up. One way to reduce this

when SW3 is set to position 2 (normal selectivity on sub-bands 1 and 3). They minimize the error caused by using the same number of turns to ground at that setting. See Table 2.

2. It is desirable that the start and finish ends of the coils on the form (as mounted) should be located on that half of the coil form nearest the front panel. This will prevent taps from being located on the far side of the coil, as viewed from the front panel, thus preventing excessive lead lengths.

3. Some experimentation in tradeoffs of the requirements in 1.) and 2.) may have to be made, since they may not be fully compatible.

The diode: This design is optimized for use with a diode having an n of 1.03 and a Saturation Current (Is) of about 106 nA at 25° C., although this is not critical and other diodes can be used with very good results. See Articles #0, 4 and 16 for info on n and Is of diodes, and how to measure them. If one has a favorite diode, its effective (Is) can be changed by applying a DC bias voltage, using perhaps, the 'Diode Bias Box' described in Article #9.

An excellent diode to use in this set is the ITT FO-215 germanium unit that was made 20 or so years ago (see Article #27). NOS may be available from from Dave Schmarder at http://www.1n34a.com/catalog/index.htm and Mike Pebble at http://www.peeblesoriginals.com

Another suitable diode, the published parameters for which show an (Is) of 100 nA is a Schottky diode, the Agilent HBAT-5400. It is a surface-mount unit that was originally designed for transient suppression purposes. Measurements of many HBAT5400 diodes seem to show that there are two varieties. One type measures approximately: n=1.03 and (Is)=80 nA The other type seems to have an n of about 1.16

Coil name	Start - inches to right or left of center	Finish - inches to right or left of center
L1	0.058 R	0.81 R
L2	0.81 R	0.99 R
L3	0.99 R	1.50 R
L4	0.058 L	0.48 L
L5	0.48 L	1.50 L

After the windings are completed, one should have a coil with ten droopy wires coming from it. The next step is to tidy up the coils, adjust the 0.12" wire spacing at the start of L1 and L4 and their one-turn taper. Move any excess winding space to slightly space-wind the last several turns of L3 and L5 at their finishes (cold ends). The turns should then be sprayed with one light application of crystal clear "Krylon" acrylic lacquer, Rust-Oleum Specialty high luster lacquer coating (clear) or equivalent, to hold them in place. The ends of L1 through L5 that are to be joined to form the taps should be cut to a length of about 0.5", tinned and soldered together as shown in Fig. 6. Pigtails for wiring to the switches should now be soldered to the taps. The coil form should be mounted with its axis parallel to the front panel as shown in Fig.1, its center about 6.50" back, and centered horizontally.

Tank coil specs. for those who wish to use a different diameter coil form, axial length of total winding, wire size or wire to wire spacing:

1. The total tank inductance, measured from point A to ground, should be 250 uH with SW3 in position 1 and 62.5 uH in position 2. Optimum partitioning of turns: L1 should comprise 26.1% of the sum of the turns of L1, L2, L3, L4, and L5; L2: 6.15%; L3: 17.75%; L4: 14.5% and L5: 35.5%. To get the highest Q in sub-bands 3 and 4 it is essential that the sum of the turns of L1, L2 and L3 equals the sum of the turns of L4 and L5. The values for L1 and L2 are compromises

problem is to rotate the unit #1-unit #2 assembly about a vertical axis and attempt to null out the pickup. Another approach could be to make physically smaller coils while still maintaining (or increasing) Q by using litz wire or by adding a trap.

Measured Performance of DT4BCS at an Output Audio Power of -70 dBW (not including Audio Transformer
Loss), using the Method described in Article #11.

Freq. in kHz.	Dial Setting: C7, C8; C7, C8	-3 dB Bandwidth in kHz.	-20 dB Bandwidth in kHz.	Ratio: (-20 dB bandwidth)/(-3 dB bandwidth).	Insertion Power Loss (S21) in dB.
603	12, 45, 36, 33	6.0	21.1	3.52	4.7
813	51, 56, 57, 50	6.7	16.3	2.43	5.7
1094	48, 27, 46, 27	6.9	22.3	3.23	6.2
1474	76, 47, 71, 43	6.6	18.6	2.82	6.4

Note: The diode rectified DC voltage at the power levels used in the measurements above is 0.5 volts (Rheostat R3 set to 350k ohms). At this power level, a SPICE simulation of the detector shows a theoretical diode detector insertion power loss of about 1.4 dB.

#23 Published: 02/07/2002; Revised: 07/27/2002

ARTICLE 24

Sensitivity and selectivity issues in crystal radio sets including diode problems; measurements of the Q of variable and fixed capacitors, RF loss in slide switches and loss tangent of various dielectrics

Part A: The Issues.

The sensitivity and selectivity in a crystal radio set can be impacted by many factors, including:

* A1: Ineffective coupling of the antenna to the tank circuit.
* A2: Resistive RF losses in the tank circuit.
* A3: Inappropriate diode.
* A4: Unintentional diode damage from exposure to electrostatic discharge (ESD) or voltages higher than the minimum reverse voltage rating. This is one cause of the loss of weak-signal sensitivity over time.
* A5: Relations between antenna-ground system, diode RF input and tank loss resistance, as effecting selectivity and loaded Q.
* A6: Audio transformer loss.

A1. There is a practical minimum limit to the possible impedance transformation ratio of the series resistance of the antenna-ground circuit to the shunt value desired across the tank circuit, when the transformation means is just a series capacitor between the antenna and the top of the tank. This problem occurs when the capacitor required for the desired impedance transformation becomes so large that it causes the tank inductance to resonate below the desired frequency (See Article #22, Part 2). The solution here is to use a lower value inductance for the tank or to tap the antenna down on the tank

Fig.7 Coil mounting hardware.

To wind the tank coil, first cut five pieces of 660/46 litz as follows: For L1: 16' 9", L2: 4' 3", L3: 11' 10", L4: 9' 11" and L5: 22' 5". These lengths provide about 9" of free length at each end. The ends of each wire length should be tinned to prevent unraveling of the strands and serving while winding the coils. Use a small Weller WC-100 adjustable temperature iron (or equivalent), set to maximum temperature (really hot!) for tinning. The method is to immobilize an end of the wire end by extending it about 2" over the end of a table and placing a weight on top of it. One can then apply the very hot iron tip to the cut strand ends and feed in a little solder to wet them. As the heat percolates down the litz and the insulation burns off or melts, more solder can be applied a little further down and around the end to obtain a solid tinning for a lead length of about 0.25". If one don't want an inch or so of litz beyond the solder to become stiffened from melted insulation, one might do as John Davidson has suggested. He tightly wraps a length of aluminum foil around the litz, up to 0.375" from the end before soldering to act as a heat sink and keep the insulation cool. I haven't tried this out.

The start leads of L1 and L4 should be spaced 0.06", right and left from the longitudinal center of the coil form. One can use a bent up plastic soda straw in the 0.25" holes to keep the wires approximately in place). Gradually taper the wires further apart going through the first turn, then to become close wound for the remainder of the windings.

Table 1 - Longitudinal locations for the taps and start/finish of L1, L2. L3, L4 and L5.

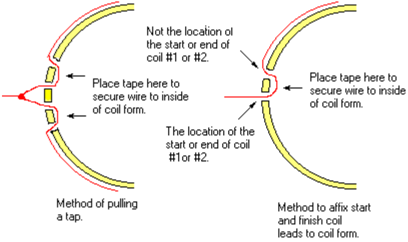
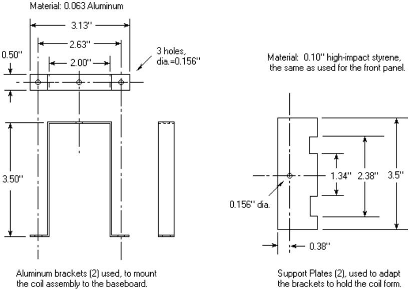


Fig. 6 - Method of forming taps and terminating windings.

Coil windings terminations



Coil mounting hardware

(towards ground). This is why some experimenters find that low inductance tank circuits seem to work better than those of higher inductance. If one does not use a series capacitor for impedance transformation, the antenna may be just tapped down on the tank. Another alternative is to connect the antenna-ground system to a low value unturned inductance that is coupled to the tank.

The main advantage of using a series capacitor connected to the top of the tank (for impedance matching) is that it moves the undesired short wave resonance of the antenna-ground circuit (present in every single tuned crystal radio set) to the highest possible frequency and reduces its strength. A disadvantage is that unless a high enough Q variable capacitor is used, insertion power loss is increased, especially at the high frequency end of the band. See Part B, Section 1 of this Article and Article #22, Part 7.

A2. Resistive RF losses in the tank circuit are affected by: 1) Losses in any capacitor used for tuning or RF coupling. 2) Physical size of the coil and such items as length/diameter ratio, cross-section size and shape, and turns spacing (to reduce coil proximity losses). 3) Loss tangent in the coil form material, wire insulation and all dielectric material penetrated by the electric field of the coil. 4) Wire size and plating, if any. Silver plating is good but tin plating is bad, especially at the high end of the band. 5) Wire construction such as litz, solid or un-insulated stranded. The latter should be avoided. 6) Switches (if used). 7) Magnetic coupling from the coil to nearby lossy metallic objects. 9) Capacity coupling from "hot" high impedance RF points through a lossy RF return path to ground. See Article #22, point 7 of Part 10. Some comments: The loss from the loss tangent of the dielectric material used for the mounting base of detector stands can be nontrivial.

Loss tangent of the material used for a front panel can cause dissipative loss if terminals provided for the connection of an external diode are too close together. See Part B, section 3 of this Article.

A3. A diode with too low an axis-crossing resistance (too high a saturation current) will resistively load the tank too heavily, causing a low loaded Q that results in loss of selectivity and sensitivity. A diode with too high an axis-crossing resistance (too low a saturation current) will increase selectivity because it only lightly resistively loads the tank circuit. The disadvantage is that sensitivity is reduced (many people liken this to the diode having too high a "turn-on" voltage.) and considerable audio distortion is generated. A little reverse or forward DC voltage bias will usually fix up these performance problems. See Article #9.

A way to check whether weak signal performance would be improved if one used a diode with a different saturation current (Is), but without experimenting with DC bias, is as follows: 1) Give the diode a one second or so spray with an aerosol "component cooler" . The reduction in temperature will temporarily substantially reduce the Is of the diode. If performance improves during the subsequent warm-back-up period, but before reaching room temperature, the diode has too high a room temperature Is. 2) Heat the diode by holding a hot soldering iron next to it for 5 seconds or so or give it a quick blow from a hot hair dryer. If performance improves during the subsequent cool-back-down period, but before reaching room temperature, the diode has too low a room temperature Is. Is, for the usual Schottky diode, changes by about two times for each 10° C. temperature change. Germanium diodes probably act the same. Aerosol component

distributed capacitance would be low in the series connection (about 7.7 pF) and higher in the parallel connection (about 21 pF), mainly because the finish (ground end) of component inductor #L(1,2,3) is located close to the start (hot end) of component inductor #L(4,5). This reduces the Q at the high end of the BC band. If the coils are contra wound, as I call it, the lower distributed capacitance condition occurs in the parallel, not the series connection, resulting a Q increase of approximately 17% at 943 kHz. It is increased even more at the high end of the BC band. The coil form is made of high-impact styrene. OD=4.5", ID=4.22" and length=3.625". I used part #S40140 purchased from the Genova Products factory retail store. (http://genovaproducts.com/factory.htm). A PVC form can be used, but its dielectric has about 4-5 times the loss of the styrene form and will reduce Q, especially at the high end of the BC band.

The start (hot) ends of L1 and L4 are affixed to the form by being lead through two 0.25" diameter holes placed 0.5" apart, measured in the circumferential direction, and held in place by 0.5" wide film tape on the inside of the form. The finish (cold) ends of L3 and L5 are affixed to the form by being lead through two 0.125" holes placed 0.5" apart, measured in the circumferential direction, and held in place as above. Three 0.125" diameter holes spaced 0.5" apart are used when pulling a tap. In Fig. 6, the red line represents the litz wire and the yellow arcs represent a cross-section of the coil form through the center of the 0.125 or 0.25" holes.

The 'contra wound' tank inductor:

L1, L2, L3 L4 and L5 are all components of the tank. It is wound with litz wire having 660 strands of #46 conductor. L1 has 13 turns, L2 has 3.125 and L3 has 8.875, all close wound (because the form is not long enough to enable spacing and the extra turns that would be required to maintain the inductance) as component inductor #L(1,2,3) with two taps. L4 has 7.25 turns and L5 has 17.75, both close wound as component inductor #L(4,5) with one tap.

The start of L1 of component inductor #L(1,2,3) is spaced 0.06" to the right of the center of the coil form, looking at the crystal radio set from the front and winding continues through L2 and L3 clockwise to about 0.25" of the right end of the form, looking at the crystal radio set from the right. The start of L4 of component inductor #L(4,5) is spaced 0.06" to the left of the center of the form, looking at the crystal radio set from the front and winding continues through L5 clockwise to about 0.25" of the left end of the form, looking at the crystal radio set from the right. When one looks at a completed contra wound inductor, one can see that the component inductors #L(1,2,3) and #L(4,5) are wound in opposite directions. Note: All coil dimensions are measured from turn center to turn center. See Article #0, Part 12 for a mini-Article about the purpose and use of the 'contra wound' inductor. Figures 2 and 3 in Article #29 provide more info on the contra wound inductor approach.

Here is the reason for this winding scheme: One can see from Fig. 4 that this crystal radio set design connects component inductors #L(1,2,3) and #L(4,5) in series for the lower half of the BC band and in parallel for the upper half. If the two coils were wound in the same direction from the hot to the cold end, as was done in the crystal radio set described in Article #22,

coolers are available from most Electronics Distributors, including Radio Shack.

A4. Semiconductor diodes are subject to damage from exposure to electrostatic discharge (ESD) and voltages higher than their reverse breakdown values. Easy-to-check-for damage shows up as an increased reverse current when tested with a reverse DC voltage close to or less than the specified minimum breakdown voltage, as compared to an undamaged diode. Other types of damage can also occur. The effect on detector performance can be anywhere from a mild to a very great reduction of weak-signal sensitivity. This type of damage has the effect of placing a resistor across the diode, which reduces tank Q and adds a parasitic, unneeded audio load resistance. If one has an old VOM having a d'Arsonval moving-coil meter movement (not a digital type) such as the Triplett 360, Weston 980 or Simpson 260, one can do a quick check of the back leakage of a Schottky diode by measuring its back resistance on the X1,000 range. If the needle does not move, the diode has probably not been damaged from ESD. This test does not apply to so-called zero-bias Schottky detector diodes because of their usual low reverse breakdown voltages. Undamaged germanium diodes normally have greater reverse leakage (lower back-resistance) than the usual Schottky detector diodes. This does not effect performance in crystal set applications. The germaniums seem less subject to the electrostatic damage problem precisely because their lower back-resistance (and high reverse breakdown voltages) tends to bleed off any electrostatic charge that might accumulate, maybe from handling. Diodes with the lowest leakage can be selected with a simple test as follows: Connect a 3-4.5 volt DC source, a 4.7k to 10k resistor, the diode and a DVM set to read DC current in series. Polarize the battery so that the diode is back biased. If the current is 2 uA or less, the parasitic back leakage resistance is greater than 2 Megohms or so and all should be well, as far as weak signal loss is concerned. The resistor is used to prevent damage to the test diode if it is accidentally connected in the forward direction. This test may be used to sort out the best weak-signal sensitivity diodes and exclude damaged ones. It is probable that for a selection of diodes having the same part number, ones with the lowest reverse leakage current will deliver the best weak-signal performance.

I believe poor weak-signal results some people have reported with the Avago (formerly Agilent) 5082-2835 and HSMS-2820 diodes can be laid at the doorstep of ESD damage. I experienced my first problem with these diodes by storing a few new ones in a propylene vitamin pill bottle for a while. When I wanted a new one I picked up the bottle, shook it (I think,) to see if something was inside and withdrew a diode. It performed badly, so I checked its back resistance with my Triplett 360 VOM as suggested above and found a low back resistance reading. Checking all the others, I found they were all bad. Static electricity did them in... beware! It was pretty stupid of me to not use anti-static bags. BTW, I've never had problems with those diodes if stored in paper bags. The bags must have had enough moisture content to provide sufficient conductivity to bleed off any static charge that might accumulate.

It is difficult to properly impedance match these diodes (see Part 4 of Article #5) because of their low saturation current (high axis-crossing resistance). They work best if 3 or or so are connected in parallel to raise the effective saturation current. If one wishes to use only one of these diodes, it could have a low forward bias applied to get the same result (see Article #9). Keep in mind that these diodes do not do very well on strong

end of the BC band, especially, sub-band 4 when the "narrow selectivity" setting is used. Performance reduction is much less in the lower sub-bands and in the "normal selectivity setting.

* Mini air-variable 365 pF caps sold by many distributors such as The Crystal Set Society and Antique Electronic Supply may be used in place of the ceramic insulated ones specified for C7 and C8. Their maximum capacitance of 365 pF may not be large enough to enable achieving the design bandwidths at the lower end of bands A and B, especially when short antennas are used. This problem can be solved by making provision for switching a 220 pF NPO (sometimes called COG) disc capacitor across C7 and C8 when tuning to these frequencies.

* Radio Shack mini DPDT switches from the 275-327B assortment or standard sized Switchcraft 46206LR switches may be used in place of the Switchcraft 56206L1 specified for SW5 and cost much less. See Article #24 for comparison with other switches.

* Molded plastic insulated rotary switches may be used for SW3 and SW4, such as those made by Lorin and sold by Mouser and others.

Wiring the crystal radio set: The stator and rotor terminals of C8 are labeled points A and B (see Fig. 5), and all connections to them should be short and direct. The purpose is to minimize spurious FM and short-wave resonances which might be created otherwise. This approach eliminates as much wiring inductance associated with C8 as possible, maximizing its ability to shunt out any high frequency spurious responses that might be present. Some more info on this subject is presented near the end of part 8.

If 300 ohm sound powered phones having their 600 ohm elements connected in parallel are used, C11 should be about 4 uF and a different transformer configuration should be used.

* **L1, L2, L3, L4, and L5: See "The tank inductor" below.

* **SW1, SW2 and SW6: DPDT general purpose slide switches. Radio Shack mini DPDT switches from the #275-327B assortment or Switchcraft #46206LR are relatively low loss units.

* **SW3 (Used to switch between bands A and B): 5 position two pole ceramic insulated with silver plated contacts rotary switch, used as a 2 position 2 pole switch. For low loss, it is essential that the switch use ceramic insulation.

* **SW4 (Used to select a sub-band and switch between normal and sharp selectivity): 5 position single pole ceramic insulated rotary switch. For low loss, it is essential that the switch use ceramic insulation.

* **SW5: Switchcraft #56206L1 DPDT mini Slide switch. Used as a SPDT switch. This switch has unusually low contact resistance and dielectric loss, but is expensive. Other slide switches can be used, but may cause some reduction of tank Q.

* T1, T2: Calrad #45-700 audio transformers. Available from Ocean State Electronics, as well as others. If 300 ohm phones are to be used, see "Audio impedance transformation", below.

* R3 (used to adjust the resistive load on the diode): 1 Meg Pot., preferably having a log taper.

* Baseboard: 12" wide x 11 1/8 " deep x 3/4" thick.

* Front panel is made of 0.1" thick high-impact styrene. Other materials may be used. This is the the lowest loss practical material I could obtain.

** For lower cost, the following component substitutions may be made: They cause a reduction in performance at the high

signals when optimally impedance matched for weak signal reception because of their low reverse voltage breakdown rating. Avago specifies that rating as 15 volts for these diodes. If a rectified voltage approaching 7.5 volts appears across the DC load (now operating in 'peak-detector' mode), a maximum reverse voltage of about 15 volts will appear across the diode for every cycle of the RF carrier. When the diode is operated at this signal level or greater, a 'bucking' rectified current flows through the diode reducing audio output and causing audio distortion. This problem is eliminated by using diodes having a greater reverse breakdown voltage such as most ITT FO 215 and most 1N34A units.

A5. A question asked by many designers is this: What is the best approach when deciding how to 'impedance match' the transformed antenna-ground-system resistance to the RF input resistance of the diode detector? This would be a no-brainer if the resonant resistance of the tank was infinite, but it is not.

Table 1 - Definitions of Terms

IPL	Insertion power loss
Lo	Inductance of the tank
Ql	Loaded Q of the tank
Qo	Unloaded Q of the tank
Rag	Actual antenna-ground-system resistance
Ragt	A antenna-ground-system resistance transformed its value at the top of the tank
Rd	Actual input RF resistance of the diode.
Rdt	Input RF resistance of the diode as transformed to its value at the top of the tank.
Ro	Resistor representing all losses in the tank
Xa	Antenna-ground-system impedance transformation.
Xd	Diode input RF resistance transformation.

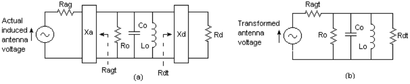


Fig. 1 - Simplified schematics of a single-tuned crystal set.

Simplified Schematic

The presence of finite tank loss increases IPL from the theoretical zero of the matched impedance case. It also reduces selectivity. Fig. 1a shows a simplified schematic of a single-tuned crystal radio set. The input impedance transformation Xa might take the form of capacitor in series with the 'Rag' and the 'induced voltage source' series combination or a tap on the tank. The diode input RF resistance transformation might take the form of a tap on the tank or a capacitor in series with the diode, along with other components. See Son of Hobbydyne and Hobbydyne II at http://www.hobbytech.com/crystalradio/crystalradio.htm.
Table 2 shows calculated data for the simplified schematic shown in Fig.1b.

Table 2 - Calculated Insertion power loss (IPL), ratio of loaded to unloaded Q and impedance match (return loss*) in a single-tuned circuit.

Parts List - All components are chosen for the best possible sensitivity at a -3 dB RF bandwidth of 5-6 kHz.

* C1, C3: 200 pF NPO ceramic caps.
* C2: 100 pF NPO ceramic cap.
* C4, C6: 270 pF ceramic caps.
* C5: 18 pF NPO ceramic cap.
* ** C7 (Antenna cap.), C8 (Tank cap.): 12-475 pF single section variable capacitors, such as those that were mfg. by Radio Condenser Corp. (later acquired by TRW). They use ceramic stator insulators and the plates are silver plated. Purchased from Fair Radio Sales Co. as part #C123/URM25. Other capacitors may be used, but those with phenolic stator insulators will cause some reduction of tank Q, especially at the high end of sub-band 4. The variable capacitors are fitted with 8:1 ratio vernier dials calibrated 0-100. These are available from Ocean State Electronics as well as others. An insulating shaft coupler is used on C7 to eliminate hand-capacity effects. It is essential, for maximum sensitivity and selectivity, to mount C7 in such a way that stray capacity from its stator to ground is minimized. See Part 11 for info on mounting C7. The variable capacitors used in this design may not be available now. Most any other capacitor with ceramic insulation should do well. Note: If one has available two capacitors having different losses for use as C7 and C8, the best one should be used for C8 since the sensitivity of this crystal radio set is less affected by a Q reduction of C7 than of C8.
* C9: 47 pF ceramic cap.
* C10: 0.1 to 0.22 uF cap.
* C11: Approx. 1.0 uF non-polarized cap. This is a good value when using RCA, Western Electric or U. S. Instruments sound powered phones with their 600 ohm elements connected in series. The best value should be determined by experiment.

sub-bands 1 and 3 and 500k ohms for sub-bands 2 and 4. Since the diode is tapped at the 0.7 voltage point for bands 2 and 4, it sees a source resistance at resonance of: 125k for sub-bands 1 and 3 and of 250k ohms for sub-bands 2 and 4. These figures apply for the theoretical case of zero loss in the tuned circuit components (infinite Q). In a shunt capacitively tuned crystal radio set, loaded with a constant resistive load, the bandwidth will vary as the square of the frequency. To understand why, consider this: When the resonant frequency of a tuned circuit loaded by fixed parallel resistance is increased (from reducing the total circuit tuning capacitance), the shunt reactance rises proportionally, giving rise to a proportionally lower circuit Q. But, a proportionally higher Q is needed if the bandwidth is to be kept constant. Therefore, the square relation.

In the practical case, we are faced with two problems. (1) How should we deal with the fact we work with finite Q components? (2) At high signal levels (above the LSLCP), the RF load presented by the diode to the tuned circuit is about 1/2 the audio load resistance, and at low signal levels (below the LSLCP) the RF load presented by the diode is about 0.026*n/Is ohms. Compromises are called for.

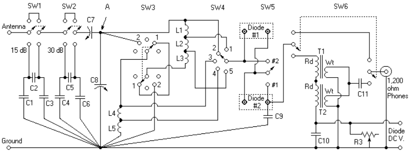


Fig. 5 - Schematic of Single-Tuned 4 Band Crystal Set (ver. C)

Schematic of version 'C'

Line #	Ragt/Ro	Rdt/Ro	IPL in dB	Ql/Qo	Input match: S11 in dB	Output match: S22 in dB
1	0.250	0.250	1.023	0.1111	-19.85	-19.85
2	0.333	0.500	1.761	0.1667	-infinity	-9.53
3	0.667	0.667	2.50	0.250	-12.04	-12.04
4	1.000	0.500	3.01	0.250	-6.02	-infinity
5	0.500	1.000	3.01	0.250	-infinity	-6.02
6	1.000	1.000	3.52	0.333	-9.54	-9.54
7	2.000	0.667	4.77	0.333	-3.52	-infinity
8	0.667	2.000	4.77	0.333	-infinity	-3.52
9	2.000	2.000	6.02	0.500	-6.02	-6.02
10	3.000	3.000	7.96	0.600	-4.34	-4.34

*Return loss is a measure of the "goodness" of an impedance match. A value of minus infinity indicates a 'perfect impedance match' (all the available input power is delivered to the load). A value of zero indicates a 'perfect impedance mismatch' (all of the available input power is reflected back to the source, and none is delivered to the load). An intermediate value indicates the amount of available power that is reflected back to the source by the load. The usual way of referring to mismatch of a two-port network is by using S parameters. S11 is the input reflection coefficient and S22, the output reflection coefficient. The magnitude of a voltage reflection coefficient is 20*log|{(Rload-Rsource)/(Rload+Rsource)}|. One reference is Radiotron Designer's Handbook, Fourth Edition, pp 891-892. In our case, consider out little circuit (at resonance) to be a zero length transmission line having attenuation. Note that the Handbook was written before 'S" parameters were widely used.

Table 1 shows the tradeoff between IPL and selectivity. The lower Rag and Rd become, the lower IPL becomes, but Ql/Qo drops (poorer selectivity). Higher values of Rag and Rd result in greater IPL and greater Ql/Qo (greater selectivity). Ramon

Vargas has suggested that many people consider the parameters on line 4 to be close to a practical optimum, and they are. Line 3 shows alternate parameters for achieving the same selectivity at an IPL 0.51 dB less. A general rule may be stated that for any given value of Ql/Qo, equal values for Rag and Rd will result in the least possible IPL. In this case, input and output return losses will be equal. It appears to me that the parameters in lines 3 or 6 are probably the ones to shoot for in most design calculations.

An effect I have observed is that one cannot simultaneously attain a perfect impedance match at both input and output ports in systems of the type shown in Fig. 1a (simultaneous conjugate match). As shown on lines 4 and 5 in Table 2, Ragt and Rdt can be arranged to provide a perfect match at one port, but then the other port will be mismatched. Ro could be replaced by a series resistor (not a real-world crystal radio set anymore) and one would still not be able to arrange a simultaneous impedance match. If circuit losses were represented by proper values of both series and shunt resistances, it would then be possible to attain a simultaneous perfect match at both input and output.

An audio transformer is one passive device that has series and shunt loss components. If it is so designed, it can provide a simultaneous perfect impedance match at both input and output when loaded by its designed-for load resistances (ignoring reactance effects). If operated at any other impedance level, such as doubling the source and load resistances, simultaneous perfect input and output impedance matching cannot be attained. This info is of mainly theoretical interest for most crystal radio set applications except when one tries to operate an audio transformer at considerably higher or lower source and load resistance values than it was designed

approaching the first resonance of the antenna-ground system. The design approach is to place a variable capacitor C3 in series with the antenna circuit (Fig. 3a) to enable impedance transformation of the antenna-ground circuit to an equivalent parallel RC (Fig. 3b), the R component of which can be adjusted by changing the value of the C3 to follow a desired relationship vs frequency. One of the objectives of the design is to enable as constant a bandwidth as possible vs. frequency. This requires the aforementioned equivalent parallel component (R2) to vary proportionally with the square of the frequency if capacitive tuning is used in each sub-band (loaded Q must be proportional to frequency for a constant bandwidth). The shunt variable capacitor to ground, shown across the tank coil, is used to tune the tank to resonance. This design attempts to accomplish this in the center of each sub-band. Performance is close at the band edges.

3. The single tuned crystal radio set

The topology of the single tuned circuit is changed from band to band as shown in Fig. 4 below.

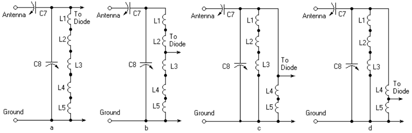


Fig. 4 - (a) Circuit topology for sub-band 1. (b) Circuit topology for sub-band 2. (c) Circuit topology for sub-band 3. (d) Circuit topology for sub-band 4. The diode connections shown are for the "normal selectivity" setting.

Sub-band circuit topology

Resonant RF resistance values at the top of C8 (Fig. 4) from antenna loading are designed to be: 250k ohms at the center of equal to each other by raising the loading resistance of the antenna-ground system and the diode on the tank by a factor of two in sub-bands 2 and 4, compared to the value used in sub-bands 1 and 3.

2. Design Approach for the Center of each of the four sub-bands.

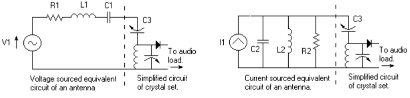


Fig. 3a
Fig. 3b

Simplified Schematic of Crystal Set

Fig. 3a shows the simplified Standard Dummy Antenna circuit, described in Terman's Radio Engineer's Handbook for simulating a typical open-wire outdoor antenna-ground system in the AM band. R1=25 ohms, C1=200 pF and L1=20 uH. See Article #20 for info on how to measure the resistance and capacitance of an antenna-ground system. These values are used in the design of the crystal radio set. R1 represents the antenna-ground system resistance, C1 the capacitance of the horizontal wire and lead-in to ground and L1 represents the series inductance of the antenna-ground system.

The values of R1, C1 and L1 in Fig. 3a are considered to be independent of frequency. To the extent that they may vary with frequency, C7 and C8 in Fig. 4 can be adjusted to compensate. The current-source equivalent circuit of the antenna-ground circuit is shown in Fig. 3b. To a first degree of approximation, C2 in Fig. 3b is independent of frequency. R2 will vary approximately inversely with frequency. We will ignore the effect of L2, since its value is large, except when

for. If this is done, IPL is considerable increased compared to using it the impedances for which it was designed.

A6. Audio transformer loss. See Articles #1 and #5.

Part B: The Measurements.

Loss measurements at various frequencies on some components, with the loss expressed as a parallel resistance (Rp) in parallel with the capacitance of the component or the loss tangent, or equivalent Q of the component:

*

B1: Q and equivalent parallel loss resistance (Rp) of two variable capacitors vs frequency when resonated with a 250 uH inductor, in parallel with a total circuit stray circuit capacitance of 20 pF. Rp and Q vs. frequency at four different capacitance settings is also shown.

*

B2: Equivalent parallel loss resistance of some DPDT slide switches.

*

B3: Loss tangent of some coil form and sheet plastic (front panel) materials.

*

B4: Q of inductors and L/C resonators.

*

B5: High Q fixed value ceramic capacitors.

B1: Measurements at 520, 730, 943, 1300 and 1710 kHz, made on two different variable capacitors used in crystal radio sets, are shown in the two graphs below. Capacitor A is a 485 pF variable capacitor that was purchased from Fair Radio Sales. It has a ceramic insulated stator and silver plated brass plates with silver plated wiper contacts. Capacitor B is a small 365 pF air-variable capacitor purchased from the Xtal Set Society. Its plates are made of aluminum and the stator support insulators are made of a phenolic plastic. This capacitor is similar to those sold by Antique Electronic Supply and others. See note at the end of this section B1.

Fig. 2 shows Rp plotted against frequency, with the capacitor adjusted at each frequency to a value 20 pf lower than that required to resonate a 250 uH inductor. This allows for a stray capacitance of 20 pF, in an actual circuit. Any losses that may be in the stray capacitance are assumed assigned to the inductor. The plot shows how, in actual practice, the Rp of variable capacitors A and B vary when tuned across the broadcast band. Fig. 3 shows how the Q of the total capacitance (including the 20 pF stray capacitance) varies across the BC band for each capacitor. Do not make the mistake, when looking at the two graphs below, of thinking that they represent Rp and Q of capacitors A or B vs. frequency, with the capacitor set to a fixed capacitance. The capacitance is varied as a function of frequency, along the horizontal axis of the graphs, to a value that would resonate with a 250 uH inductor. Figs. 4 and 5 show Rp and Q of capacitor B as a function of frequency, at four different fixed capacity settings (the frequency at each capacitance setting is always a value that brings about resonance with the 250 uH inductor).

varying CT by 11:1 will cover the range, but have two disadvantages. (1) The -3dB bandwidth will vary by 1:11 from 520 to 1740 kHz. (2) In the practical case, if the bandwidth is set to 6 kHz at the low end of the BC band, and an attempt is made to narrow the bandwidth at 1710 kHz by placing a capacitor in series with the antenna, the insertion power loss will become great.

3. The compromise used here is a coil design that can be switched between two inductance values differing by 4:1. The high inductance setting is used for the low half of the BC band and the low inductance for the high half. Capacitive tuning is used to tune across each half. The new technique used here, of using a combination of two inductors, enables the Q of the low value inductance (used in the high half of the band) to to be much higher than would be the case if a single coil of the same diameter and wire size, but with fewer turns, were used. This technique uses two coils, closely coupled, and on the same axis. They are connected in series to obtain the large inductance and in parallel for realization of the small one. The small inductance has a value 1/4 that of the large one and about the same Q at 1 MHz. (if coil distributed capacity is disregarded). The innovation, so far as I know, is to use the full length of wire used in the high inductance coil, occupy the same cubic volume, but get 1/4 the inductance and keep the same Q as the high inductance coil (at the same frequency). See Table 4.

4. The high and low bands are each further subdivided giving a total of four sub-bands (1, 2, 3 and 4). If this were not done, we would be faced with a bandwidth variation of about 1:3.3 in each band. The geometrically subdivided bands are: sub-band 1 (520-700 kHz), 2 (700-943 kHz), 3 (943-1270 kHz) and 4 (1270-1710 kHz). The bandwidth should vary about 1:1.8 across each of these sub-bands. The bandwidths at the center of each of the four sub-bands are made approximately

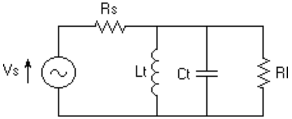


Fig.2 - Single-tuned circuit

Basic single tuned schematic

The frequency response shape of the circuit shown in Fig. 2 is that of a simple single tuned circuit and can be thought of as representative of the nominal response of a single tuned crystal radio set. Consider these facts:

1. If Lt and Ct have no loss (infinite Q), zero insertion power loss occurs at resonance when Rs equals Rl. This is called an 'impedance matched' condition. The power source (Vs, Rs), looking towards the tank, sees a resistance value equal to itself (Rl). Also, the load (Rl), looking towards the input sees a resistance (Rs), equal to itself. In the practical case there is a finite loss in Lt and CT This loss can be represented by an additional resistance Rt (not shown), shunted across the tuned circuit. The input resistance seen by (Vs, Rs) is now the parallel combo of Rt and Rl and it is less than Rs The perfect impedance match seen by (Vs, Rs) when the tank Q (Qt) was infinite is now destroyed. The impedance matched condition can be restored by placing an impedance transformation device between the source, (Vs, Rs) and the tank.

2. In Fig. 2, if tuning could be done with Lt alone, leaving CT fixed, the bandwidth would be constant. The problem here is that high Q variable inductors that can be varied over an approximately 11:1 range, as would be needed to tune from 520 to 1710 kHz do not exist. On the other hand, tuning by

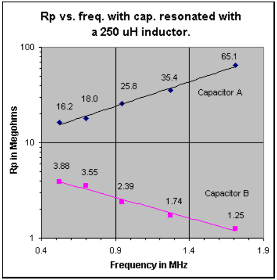


Fig. 2

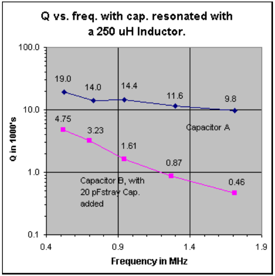


Fig. 3

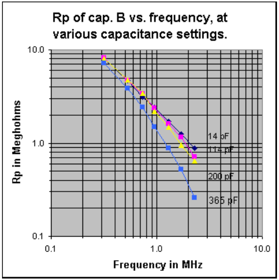


Fig.4

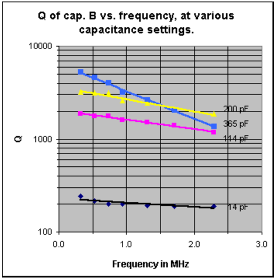


Fig. 5

addition of the amplifier in Article #25. This Article does not include the amplifier.

Design objectives:

* A relatively constant -3 dB bandwidth of 5 to 6 kHz across the full range of 520 to 1710 kHz at normal selectivity, with a relatively constant RF power loss in the RF tuned circuits of less than 3 dB.

* A switched adjustment to achieve about 3 times sharper selectivity than normal.

* Optimal performance with external antenna-ground systems having a fairly wide range of impedance.

* To provide a simple-to-use switching setup for comparing a 'test' diode with a 'standard' one.

* To provide a volume control with a range of 45 dB in 15 dB steps that has the minimal possible effect on tuning, This was incorporated in the design because the two local 50 kW blowtorch stations WABC and WOR (about 10 miles away) deliver a very uncomfortably loud output from SP headphones from my attic antenna. A means of volume reduction that did not reduce selectivity was needed. This method of volume reduction actually increases selectivity by isolating antenna-ground resistance from the tank circuit.

* Introduction fo a new (to me) method for constructing low inductance high Q coils.

1. Theory

4 at point 'A'. The two different transformed RF antenna-ground system resistance values are achieved by proper adjustment of a variable capacitor in series with the antenna (C7 in Fig. 5). The higher diode RF tank loading resistance value for sub-bands 2 and 4 are achieved by tapping the diode onto the tank at a point that is 70% of the turns up from ground. The tank is not tapped for sub-bands 1 and 3 and connection is to the top of the tank. In the sharp selectivity mode the diode is tapped half of the turns down on the tank from the point used for normal selectivity.

* The weak-signal RF input and audio output resistances of a diode detector are approximately the same and equal to 0.026*n/Is ohms (Is means diode saturation current, see Article #0-Part 4). The strong-signal audio output resistance of a diode detector is approximately equal to 2 times the RF resistance of its source. Compromise audio impedance transformation ratios are used to optimize performance on both weak and strong signals.

* The design is scalable. Less expensive parts that may have somewhat greater losses may be used with some penalty in sensitivity and selectivity, especially at the at the high end of the BC band and at the Sharp Selectivity setting. See the Parts List for a listing of some more easily available and lower cost parts than the ones used in the original design.

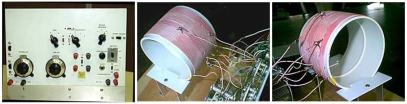


Fig. 1 - Single-Tuned Four-Band Crystal Radio Set, Version 'C'. These are actually pictures of Version B as described in Article #22, converted to version C, as described in this Article, but modified with the

Discussion:

There are two main sources of loss in an air-dielectric variable cap: 1) Loss in the dielectric of the stator insulators, and 2) Resistive losses in the metal parts. Of course, there is also the important very low-loss capacitor made from the air dielectric between the plates. The losses in 2) include resistive loss in the plates and in the wiper that connects the rotor to the frame. Resistive loss in the plates is very small at low frequencies, but increases with increasing frequency because of skin effect.

Notice, that for capacitor B, Rp (vertical scale) in Fig. 4 is about the same for all capacitance settings at the lowest frequency plotted (310 kHz). Rp is approximately constant over the full 14-365 pF capacitance range. This is because we are varying a high Q air dielectric capacitor of very low loss, in parallel with a low Q fixed capacitance made up of the lossy phenolic stator supports. Most of Rp comes from the loss in the phenolic. In Fig. 5, again at the low frequency end of the horizontal axis, observe that Q is a direct linear function of the overall capacitance, as it should be. The Q of the air-variable part, taken alone, is much higher than that of the capacitor made up of the phenolic insulators. The main loss, here, comes from the approximately fixed shunt Rp provided by the phenolic stator supports. At 14 pF the C from the air cap is relatively low compared to that from the phenolic insulators. At 365 pF the C from the air cap is much higher than that from the phenolic supports.

Things change at higher frequencies. The reactance of the capacitor drops. This, combined with the series resistance of the plates and wiper now come into play as an additional factor reducing the Q. If this series resistance (Rs) were the only resistance affecting Q, the equation for Q would be: Q=(reactance of the capacitor)/Rs. One can see from this equation that the introduction of Rs makes the Q drop when frequency increases. Up to now, the main loss came from the parallel loss resistance of the phenolic supports. The Q from a setting of 365 pF drops quite rapidly with increase of frequency because of this series resistance. Skin effect makes the effect worse by increasing Rs, the higher one goes in frequency. Notice that at the low frequency end of Fig. 5, Q is approximately proportional to the capacitance, as it should be if the main loss is the fixed shunt resistance Rp, coming from the phenolic stator insulators.

Fig. 5 shows an approximately constant Q (vs. frequency) when the capacitor is set to 14 pF. The loss causing this low, constant Q, comes from the loss tangent of the capacitor formed from the phenolic stator insulators. At low frequencies, Q increases when the capacitor is successively set to 114, 200 or 365 pF because engaging the plates adds a high Q air dielectric capacitor component in parallel with the low Q capacitor formed from the phenolic dielectric supports. As frequency increases, with the capacitor set to 365 pF, one can see that the Q drops at a faster rate than it does when set to 200 or 114 pF. This is because Rs (being in series with the air capacitor, that dominates at the 365 pF setting), acting with its lower capacitive reactance results in a capacitor of lower Q (Q=reactance of air capacitor at 365 pF/Rs).

In the Figs. 2 and 3, the capacitor is always set for a circuit capacity value that resonates with 250 uH. This means that at 520 kHz, the varicap is set so the circuit capacity is 375 pF. At 1710 kHz, the circuit capacity is set to 34.7 pF. Even though the capacitor Q goes from 19,000 to 9800 as frequency goes from 520 to 1710 kHz, Rp increases as frequency increases

immunity from the Q reducing effects of surrounding high loss dielectric materials such as baseboard etc. The lossy stray capacity introduced is better swamped out by the high shunt tuning capacity used.

The Crystal Radio Set Design, in a (large) Nutshell:

* The design approach is to divide the AM band into several sub-bands in an attempt to keep the selectivity relatively constant and the insertion power loss low across the whole band.

* The first step is to divide the BC band into two halves: band A (520-943 kHz) and band B (943-1710 kHz). Two-step shunt inductive tuning is employed to switch between bands. A tank inductance of 250 uH is used in band A and 62.5 in and B.

* Band A is further subdivided into two bands: sub-bands 1 (520-700 kHz) and 2 (700-943 kHz). The band B is also subdivided into two bands: sub-band 3 (943-1270 kHz) and 4 (1270-1710 kHz).

* In the normal selectivity mode, two different resonant RF resistance levels, measured at the top of the tuned circuit (point 'A' in Fig. 5), are used at the center of the sub-bands. This impedance level is about 125k ohms at the center of sub-bands 1 and 3. It is 250k at the center of sub-bands 2 and 4 (excluding the resistive losses of the components used). These resistance values are made up of a parallel combination of the transformed RF antenna-ground system resistance and the input RF resistance of the diode. These two resistances should be equal to each other to achieve minimum insertion power loss, at the design bandwidth. This means that the transformed antenna-ground system and diode RF resistances are each about 250k in sub-bands 1 and 3 and 500k in sub-bands 2 and

insulated wire used on version 'B'. A new way to make a higher Q, low inductance coil is described.

Some additional benefits of the "Benodyne" type of tank circuit are: (1) Reduction of the the sharp drop in tank Q or sensitivity at the high frequency end of the BC band often experienced when only one value of tank inductance is used for the whole BC band. (2) Reduction of the tank Q loss from the variable cap when using lower cost units that use phenolic insulation, such as the common 365 pF cap (see Figs. 2, 3, and 4 in Article #24). The "two inductance value benodyne* circuit is used in the crystal radio sets in Articles #22 and #26. We will assume here that the two "benodyne" component inductors (see "The Tank Inductor" in this Article) provide a tank inductance of 250 uH in the low frequency half of the BC band (520-943 kHz) and 62.5 uH in the high half (0.943-1.71 MHz). If the large 250 uH inductance setting were used all the way up to the top end of the BC band (as in the usual case), a total tuning capacity of 34.7 pF would be required at 1.71 MHz (Condition A). In the "Benodyne" circuit, with the 62.5 uH inductance setting used for the high frequency half of the BC band, a total tuning capacitance of 139 pF is required at 1.71 MHz (Condition B). Benefit (1) occurs because in condition A, a larger fraction of the total tuning capacitance comes from the typically low Q distributed capacity of the inductor than in condition B. This results in a higher Q total capacitance in condition B than in condition A. Benefit (2) occurs because the effective Q of a typical 365 pF variable cap, when used with a 250 uH tank, is about 500 at 1.71 MHz (see Fig. 3 in Article #24). The Q of the 365 pF variable cap, when set to 139 pF, is greater than 1500 (see Fig.5 in Article #24). This higher Q results in less loss and greater selectivity at the high end of the BC band in condition 2. A further benefit of the Benodyne circuit at the high end of the BC band is greater

because the circuit capacity must be reduced from 375 to 34.7 pF to tune the tank from 520 to 1710 kHz (Rp=Q/(2*pi*f*C).

If the question is posed: 'Is it more important to have ceramic insulated stators or silver plated plates on a variable capacitor used in a BC band crystal radio set?', the answer is that ceramic insulated stators are the way to go.

Capacitor A: The main practical conclusion that can be taken from Fig .2 above is that Rp of capacitor A is very high over the whole band and varies roughly proportionally to frequency, over the frequency range of interest: 520-1710 kHz In fact, Rp is so high that it will not contribute any appreciable loss even when used in high performance crystal radio sets using a high Q tank inductor. Another plus is that its Rp increases with increasing frequency, further reducing any effect on loaded Q, sensitivity or selectivity at the high end of the band. The silver plating on the brass plates is beneficial because the resistivity of silver is about 25% of that of brass. Practically speaking, the silver should have little effect on the operation of a crystal radio set in the BC band, but short wave is another matter. The Q of the capacitor drops at higher frequencies, especially when set to a high capacitance value. Silver plating can materially improve performance at higher frequencies by providing a needed higher Q. Be aware that some capacitors are made with cadmium plated brass plates. Silvery-whitish colored cadmium has a resistivity 4.6 times that of silver. This higher resistivity will somewhat reduce Q at high frequencies and at high capacity settings. Some people mistakenly assume, because of the silvery-whitish color, that cadmium plated plates are really silver plated.

Capacitor B: One can see from Fig. 2 that Rp of capacitor B varies approximately inversely with frequency. The DLF (dielectric loss factor) of the phenolic stator support insulators is the main cause of this loss, over the whole frequency band. Towards the high end of the band, some loss is contributed by the series resistance in the capacitor plates, the rotor shaft wiper contact and skin effect. This loss effect is greater than that in capacitor A because the resistivity of the aluminum plates is 1.7 times that of silver. Practically speaking, this effect is minimized at the high end of the band because the plates are mostly disengaged. The Rp of the capacitor will have its greatest effect in reducing sensitivity and selectivity at the high frequency end of the band because that is where its value is the the lowest. See note at the end of this Article.

The usual crystal radio set uses shunt capacitor tuning with a fixed tank inductance. This configuration causes the tank reactance to be highest at the high end of the band, thus further reducing loaded Q, sensitivity and selectivity, for a given value of Rp. At the low and medium frequency parts of the band, Rp of the capacitor is so high that its effect is small in many crystal radio sets. Highest performance crystal radio sets made with high Q inductors, with careful attention to impedance matching may experience a noticeable reduction in sensitivity and selectivity at the high end of the band when using this or other capacitors using phenolic stator insulation. This is because the Rp of the capacitor becomes comparable to the higher equivalent Rp of the high Q tank inductor at the high end of the band as compared to its value at the low end. To clarify this, an ideal condition would exist if Rp of the capacitor and inductor were infinite. If this were to be the case, and good impedance matching of antenna-ground system to tank, diode to tank and headphones to diode existed, all of the power intercepted by the antenna-ground system would be delivered to the headphones and maximum sensitivity would occur. Any loss present in capacitors, inductors or

ARTICLE 26

Highly sensitive and selective single-tuned four-band crystal radio set using a new contra wound dual-value inductor, and having a 'sharp selectivity setting'; along with a way to measure the unloaded Q of an L/C resonator

Summary: This article describes Version 'c' of a single-tuned four-band crystal radio set, sometimes called a "Benodyne" (constant bandwidth with maximum weak-signal sensitivity across the whole BC band, achieved by using 2 or more values of inductance in the tank - lower values at higher frequencies). It is designed for a constant bandwidth across the AM band, two selectivity settings (normal and sharp) and low loss (high sensitivity) especially for weak signals at the high end of the band. It is an attempt to achieve the two objectives at a -3 dB RF bandwidth of about 5-6 kHz (relatively independent of signal strength), with a constant high efficiency across the entire AM broadcast band (normal selectivity setting): 1) Best possible sensitivity on weak signals, 2) Loudest possible volume on strong signals. The sharp selectivity setting reduces the -3dB bandwidth to about 2 kHz, but unavoidably introduces some extra insertion power loss.

Selectivity and insertion power loss figures from a computer simulation are given and compared with those of the actual physical crystal radio set. A way to tell if the detector is operating below, at or above its 'Linear-to-Square Law Crossover point' (LSLCP) is described. See Article #15a for a discussion of LSLCP. No antenna tuner is necessary for the average outdoor or attic antenna. An explanation of 'short wave ghost signals' and 'hash' is provided along with some suggestions on how to combat them. Version C uses a tank coil constructed with litz wire, as compared to the solid teflon

* Mini air-variable 365 pF caps sold by many distributors such as The Crystal Set Society and Antique Electronic Supply may be used in place of the ones specified for C7 and C8.

* 18 Ga. "bell wire" supplied by many distributors such as Home Depot, Lowes and Sears may be used in place of the teflon insulated wire specified. This vinyl insulated non-tinned copper wire is sold in New Jersey in double or triple twisted strand form for 8 and 10 cents per foot, respectively. The cost comes out as low as 3 1/3 cents per foot for one strand. The main catch is that one has to untangle and straighten the wires before using them. I have used only the white colored wire but I suppose the colored strands will work the same (re dielectric loss). The measured OD of strands from various dealers varied from 0.065 to 0.079". The extra dielectric loss factor of the vinyl, compared to the teflon will cause some reduction of sensitivity and selectivity, more at the high end of the band than the low end.

* Radio Shack mini DPDT switches from the 275-327B assortment or standard sized Switchcraft 46206LR switches work fine in place of the specified Switchcraft 56206L1 and cost much less. See Article #24 for comparison with other switches. Any switch with over 4 Megohms Rp shown in Part 2 of Article #24 should work well as far as loss is concerned. Overall, losses in the switches have only a very small effect on overall performance.

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transformers reduces sensitivity. In this discussion we are dealing primarily with losses in the resonating capacitor and tank inductor, both referred to as their respective values of Rp. The higher the value of these Rps, compared to the transformed antenna-ground source resistance across which they appear, the lesser the loss they cause. One approach to counter the effect the drop in Rp as frequency increases is to change to two-step inductive tuning by dividing the band into two sections as described in Article #22.

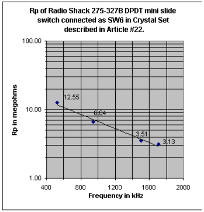
Technical Note: At any frequency, a real world capacitor of value C1 having a Q of Q1 (Q1>10), can be quite accurately modeled as a series combination of an ideal no-loss capacitor of value C1 and a resistor (Rs) equal to: (reactance of C1)/Q1. Alternatively, at any frequency, a capacitor of value C1, having a Q of Q1 (Q1>10) can be modeled as a parallel combination of an ideal capacitor of value C1 and a resistor (Rp). The resistor, Rp in this case, has a value of:: (reactance of C1)*Q1. Note that since capacitor reactance is a function of frequency, the value of the resistor will, in general, vary with frequency. In crystal radio set design it is sometimes convenient to model the tuning capacitor loss as a parallel resistor, other times as a series resistor.

Credit must go to Bill Hebbert for making the time consuming, difficult, precision measurements required for Figs. 2-5.

B2: Slide switches used in the Crystal Radio Set described in Article #22 have dielectric losses, as do all switches. To get a handle on this loss, samples of several different types of DPDT switch were measured at 1500 kHz. Each switch except the last three, below, were wired as a SPDT unit by paralleling the two sections. The Q of the capacitance appearing across the open contacts was then measured and Rp was calculated. Rp usually varies approximately inversely with frequency and therefore causes more loss at the high end of the band than at the low end. Contact resistance of all switches was found to be very low. It is unknown how well that characteristic will hold up over time. Note the extremely low loss of the Switchcraft 56206L1. The loss is so low that this switch is overkill in most crystal radio set applications. For applications in which the crystal radio set builder wants to use the lowest loss DPDT slide-switch available, this switch is the best I have found.

Table 3 - Equivalent parallel Loss Resistance (Rp) caused by the loss tangent of the dielectric of some DPDT Slide Switches at 1.5 MHz.

Brand name, model number and Size of switch	Rp in Megohms
ARK-LES (std. size)	2.3
Radio Shack 275-403A (std. size)	3.5
Radio Shack 275-407A (sub mini)	4.1
Stackpole S7022X (std. size)	4.9
Switchcraft 46206EE-6 (std. size)	5.1
Radio Shack 275-327B (mini)	5.7
CW (mini)	6.0
Switchcraft 46206LR (std. size)	6.5
C&K L202-1 (mini)	6.8
Switchcraft 56206L1 (mini)	13.3
Radio Shack 275-327B connected as SW3 in Article #22 (HI band)	4.1
Switchcraft 56206L1 connected as SW6 in Article #22	9.4
Radio Shack 275-327B connected as SW6 in Article #22, **See Graph--->>**	3.5



Rotary selector switches using ceramic insulation should have very low loss, even lower than the Switchcraft 56206L1. Quality switches using brown phenolic insulation probably have losses similar to slide switches using similar material. I

* SW1, 2, 7 and 9: DPDT general purpose slide switches.
* **SW3, 4, 5 and 6: Switchcraft #56206L1 DPDT mini Slide switches. This switch has unusually low contact resistance and dielectric loss, but is expensive. Other slide switches can be used, but may cause some small reduction of tank Q.
* SW8: 3P2T slide or other type switch.
* SW10: 3 position rotary switch.
* T1, T2: Calrad #45-700 audio transformer. Available from Ocean State Electronics, as well as others. If 300 ohm phones are to be used, see the third paragraph after Table 1.
* T3: Bogen T725 4 watt P/A transformer. Available from Lashen Electronics, Grainger or other sources (http://www.lashen.com/vendors/bogen/Speaker_Transformers.asp)
* R3: 1 Meg linear taper pot.
* R4: 10k resistor
* R6, R7: 10 Meg resistors. For minimum waveform clipping in IC1, values of R5 and R7 should be selected to be within 5% of each other.
* R8, R9: 2.2 Meg resistors. For minimum waveform clipping in IC!, values of R8 and R9 should be selected to be within 5% of each other.
* R10: 10 Meg resistor
* Baseboard: 12" wide x 11 1/8 " deep x 3/4" thick.
* Front panel: 0.125" thick high-impact styrene. Other materials can be used. I was looking for the lowest loss, practical material I could obtain.

** For lower cost, the following component substitutions may be made: Together they cause a small reduction in performance at the high end of the band (1.75 dB greater insertion power loss and 1.5 kHz greater -3 dB bandwidth). The performance reduction is less at lower frequencies.

* C13: 0.1 to 0.33 F electric double layer capacitor (supercap). Elna's 0.33 F "Dynacap", available from Mouser as #555-DX5R5H334 or Panasonic's 0.033 F "Gold" capacitor #EEC-S0HD334H, available from Digi-Key etc. are suitable. Do not use an ordinary electrolytic cap in this application. Its leakage current will probably be so great that C13 can only charge to a low voltage, and it won't be able to hold a charge anywhere near as long as a supercap. A 0.33 F supercap will charge more slowly, but it will last longer than on a 0.1 F supercap.

* C14: 1.0 nF ceramic cap

* C15: 470 nF plastic film cap. (polyester or mylar)

* C16: 10 nF ceramic cap (Connect with short leads across + and - supply voltage terminals of IC1.)

* ** L1, L2, L3 and L4: Close coupled inductors wound with uniformly spaced teflon insulated 18 Gage silver plated solid wire. This wire is used only to gain the benefit of the 0.010" thick low-loss insulation that assures that no wandering turns can become 'close-spaced'. L1 has 12 turns, L2 has 8 turns, L3 has 6 turns and L4 has 14 turns. The coil form is made of high-impact styrene. I used part #S40160 from Genova Products (http://genovaproducts.com/factory.htm). A piece of plastic drain pipe of the same OD, made of ABS, can also be used, with the same results. PVC pipe will result in somewhat less selectivity and sensitivity. See Fig. 6 for hole drilling dimensions.

* IC1: Texas Instruments micropower opamp OPA349UA (Formerly a Burr-Brown product.) Here is a link to the data sheet for this IC: http://www-s.ti.com/sc/ds/opa349.pdf A convenient way to connect to the tiny leads of IC1 is to first solder it to a surfboard such as one manufactured by Capital Advanced Technologies (http://www.capitaladvanced.com). Their models 9081 or 9082 are suitable and are available from various distributors such as Alltronics, Digi-Key, etc.

would expect that the slope of the Rp vs. frequency graph of the other slide switches to be the similar that shown above.

B3: The loss tangent of an insulating material is the reciprocal of the Q of a capacitor made of that material. Some of the insulating materials listed below are used as front panels, detector stand bases, wire insulation and coil forms in crystal radio sets. A capacitor formed by the use of one of these materials, connected across a high impedance point and ground, will contribute a loss proportional to the loss tangent and capacitance.

Table 4 - Loss Tangent of some Insulating Materials used in Crystal Radio Sets, Measured at 1.5 MHz.

Dielectric Material	Loss tangent	Q of a capacitor using the material
PVC as used in coil form	0.017	59
PVC wire insulation	0.03	33
High-impact styrene coupler from Genova Mfg. Co. (opaque white)	0.0017	590
Polypropylene 1.5" diameter drain pipe from Genova Mfg. Co.	0.0022	452
High impact styrene sheet, 0.1" thick (opaque white and flexible)	0.0023	430
Plexiglas 0.115" thick	0.016	55
FR-4 PCB material 1/16' thick	0.027	37
Black 3/16" Condensite panel, brand name "Celoron" (Bakelite, new old-stock radio panel from the '20's)	0.035 (@ 0.8 MHz)	29
ABS styrene (black)	0.010	100
Garolite (black)	0.033	30
ABS styrene (light beige)	0.020	50
HDPE (milky white)	0.0009	1120
Polystyrene (light brown opaque)	0.0032	320

Note: GE's version of polycarbonate is called Lexan. A review GE's spec. sheets of various grades of Lexan show loss tangent values at 1.0 MHz ranging between 0.006 and 0.026 . Many grades are specified 0.01.

B4: Please see Parts 10 and 11 of Article #26 as well as Table 4. Also see Table 2 of Article #22 and Article #29.

B5: Sometimes, when working with high Q tank circuits, a need pops up for a fixed capacitor with a value between say, 100 and 1000 pF that will not degrade overall circuit Q. Generic NPO disc capacitors in that range usually have a Q of around 2000-3000 at 1 MHz. Table 4 shows some caps having higher Q values. The only downside to the high Q caps is that they are SMD types and require some skill when soldering pigtail leads to them to easy connecting to one's circuit. The capacitors were measured singly, in parallel or in series, aiming for values approximating 500 pF. This was for convenience in measurement. I used solid tinned copper wire having a diameter of about 0.010" for my pig-tails. The source for the strands was a piece of stranded hook-up wire.

Table 4 - Q of easily available capacitors in the 100 - 1000 pF range

	Type	Value in pF	Voltage	Q at about 920 kHz	Mfg	Mfg. part number
1	Polypropylene	Two 1k in series=500	630	2,000	Xicon (Mouser)	1431-6102K
2	Polystyrene	One 470	50	6,200	Xicon (Mouser)	23PS147
3	Generic NPO disc	One 220	500	2,800	3/8" dia.	NA
4	Multilayer, hi Q SMD	Two 1k in series=500	50	8,000	Murata	ERB32Q5C1H102JDX1L
5	Multilayer, hi Q SMD	One 470	100	18,000	Murata	ERB32Q5C2A471JDX1L
6	Multilayer, hi Q SMD	Two 220 in parallel=440	200	20,000	Murata	ERB32Q5C2D221JDX1L
7	Multilayer, hi Q SMD	Two100 in parallel=200	500	40,000	Murata	ERB32Q5C2H101JDX1L

Note: Solder flux contamination on a dielectric is the enemy of high Q because it usually provides a resistive leakage path. If one gets solder flux on the insulation of a variable cap or switch, remove it with a commercial flux remover. This is important when a DX crystal radio set is involved.

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easily adapted to the higher performance crystal radio set described in Article #26 as well as others.

* C1, C3: 200 pF NPO ceramic caps.
* C2: 100 pF NPO ceramic cap.
* C4, C6: 270 pF NPO ceramic caps.
* C5: 18 pF NPO ceramic cap.
* ** C7, C8: 12-475 pF single section variable capacitors, such as those that were mfg. by Radio Condenser Corp. (later TRW). They use ceramic stator insulators and the plates are silver plated. Purchased from Fair Radio Sales Co. as part # C123/URM25. Other capacitors may be used, but some of those with phenolic stator insulators probably will cause some reduction of tank Q. The variable capacitors are fitted with 8:1 ratio vernier dials calibrated 0-100. These are available from Ocean State Electronics as well as others. An insulating shaft coupler is used on C7 to eliminate hand-capacity effects. It is essential, for maximum sensitivity, to mount C7 in such a way that stray capacity from its stator to ground is minimized. See Part 9 of Article #22 for info on mounting C7. The variable capacitors used in this design may not be available now. Most any other capacitor with silver plated plates and ceramic insulation should do well.
* C9: 47 pF ceramic cap.
* C10: 100 nF cap.
* C11: 1.0 uF non-polarized cap. This is a good value when using RCA, Western Electric or U. S. Instruments sound powered phones, with their 600 ohm elements connected in series. The best value should be determined by experiment. If 300 ohm sound powered phones having their 600 ohm elements connected in parallel are used, C11 should be about 4 uF, and a different transformer configuration should be used.
* C12: 0.1 uF cap

Charge/discharge considerations for C13: C13 (0.33 F) will charge close to full capacity after about 24 hours of charging. The first charge will not last as long as subsequent ones because of a phenomenon known as "dielectric absorption". If C13 is reduced to 0.1 F, about 8 hours are needed. Listening time when using headphones should be greater than 24 hours when using 0.33 F, and 10 hours when using a 0.1 F value. There is a greater current drain on C13 when using a speaker, and the listening time will depend upon the volume setting. Listening times approximate 10 hours when using a 0.33 F cap and 3 hours when using 0.1 F.

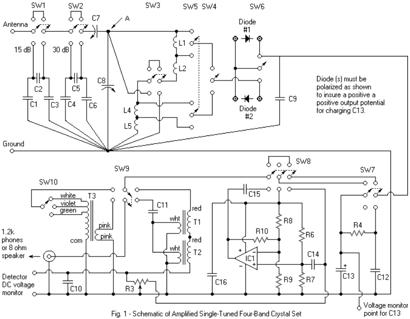

Fig. 1 - Schematic of Amplified Single-Tuned Four-Band Crystal Set

Schematic for amplified crystal set

Parts List when the crystal radio set used is the that described in Article #22. The amplifier is

ARTICLE 25

A new approach to amplifying the output of a crystal radio, using energy extracted from the RF carrier to power a micro-power IC to drive headphones or a speaker

Quick summary: This article describes an amplifier that can be used to substantially increase the volume from a crystal radio set when tuned to a weak signal when using headphones. It can also be used to amplify the output of a crystal radio set, when tuned to medium or strong stations, to drive a speaker. No battery for powering is required. The amplifier can be added to most any crystal radio set, provided access to a strong station is available. As shown here, the amplifier is applied to Version 'B' of the "Benodyne", a single-tuned four-band crystal radio set. See Article #22. It has been also applied to version ""C" of the "Benodyne", described in Article #26. A switch is provided so the crystal radio set can perform as it normally does, or with about 20 dB of audio amplification (+20 dB represents a large increase of volume.). This amplification is provided by a micropower integrated circuit that does not use battery power. Power to operate the integrated circuit is stored in an electric double-layer "supercapacitor" that can be charged overnight by leaving the crystal radio set on, tuned to a strong local station. One charge can last for tens of hours when listening to weak stations. For loudspeaker operation, a large reentrant horn type PA speaker is best, for the highest volume, although other types may be used. Depending upon volume, a full charge on the capacitor can last for about 5 hours of low volume loudspeaker listening.

The Amplifier, applied to a crystal radio set:

This crystal radio set operates in the same manner as the ones described in Articles #22 (Benodyne version B)and #26 (Benodyne version C) when switches SW7 and SW 8 are in their 'up' positions and SW9 is to the right. To operate the amplifier, first, supercapacitor C13 must be charged up to at least 1.5 volts. The manufacturer of the IC specifies a minimum of 1.8 volts, but so far, I have found that 1.5 volts to be sufficient. To charge C13, set SW7 and SW8 to their 'up' positions, SW9 to the right, and the wiper arm of R3 to the center (see Fig. 1). Tune in a station that provides between 1.3 and 5 .5 volts DC at the 'Detector bias monitor' terminals. If the voltage is too low, try changing the antenna impedance matching by optimizing the settings of C7 and C8. Set SW7 to its down position and C13 will start charging. If no station exists that is strong enough to supply at least 1.5 volts, C13 may be charged by connecting a series combination of a 4.5 volt battery and a 100 ohm current limiting resistor across it for about 30 minutes. Make sure the + side of C13 is charged positive. After C13 is charged, set SW7 to its up position. The higher the final charged voltage on C13, the higher the maximum volume will be.

Non-amplified operation with Sound Powered 1200 ohm headphones: SW7 and SW8 are up and SW9 is to the right. R3 is used to optimize DC current in the diode for minimum audio distortion.

Amplified operation with Sound Powered 1200 ohm headphones on very weak signals: SW7 is up, SW8 down and SW9 is to the right.

Amplified operation, driving an 8 of 16 ohm speaker from medium and strong stations: Operate SW7 to its up position, SW8 down and SW9 to the left. The speaker will probably

give forth with some distorted audio. To reduce the distortion, try adjusting R3. If this doesn't help enough, reduce the signal into the amplifier. The attenuators, controlled by SW1 and/or SW2 can be used to do this (see Fig. 1). If no SW1 or SW2 is present, reduce the output of the detector by decoupling the antenna (reduce C7 and restore tuning using C8).

Switch SW10 provides a tradeoff between maximum volume and current drain. Switching to the white wire connection gives the, longest listening time, but with a lower maximum volume. Each listener must make his own choice here. The current drain from C13 and the life of its charge are directly proportional to the strength of the audio signal and the setting of SW10. Maximum low-distortion volume is proportional to the voltage charge on C13.

For comparison purposes with receiving locations other than mine, there are two 50 kW stations about 10.5 miles from my home. They are WOR and WABC. My attic antenna is described in Article #20. Either station can deliver about a 5.0 volt charge to C13.

This crystal radio set was constructed by modifying a Version 'b' crystal radio set (See Article #22), using the air-mounted, flying joint method of wiring the amplifier components. A convenient way to connect to the tiny leads of IC1 is to first solder it to a surfboard such as one manufactured by Capital Advanced Technologies (http://www.capitaladvanced.com). Their models 9081 or 9082 are suitable and are available from various distributors such as Alltronics, Digi-Key, etc. The amplifier can be built in as an addition to any crystal radio set if proper allowance is made for impedance matching considerations.