KF6VSG HamPost

Software Science Inc

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Welcome to the *HamPost*

Hello, and welcome to the first edition of *HamPost*, a free quarterly newsletter by and for radio amateur operators.

The purpose of this newsletter is to provide specialized information to an audience of hams that range in experience from the newly-ticketed through veteran Amateur Extras.

Probably you already subscribe to QST, visit popular ham sites like eham.net, qrz.com or arrl.net, perhaps attend the monthly meetings of your local amateur radio club, and almost certainly have made a pilgrimage to swap meets, ham fests, and ham conventions. I do all those things too, but amazingly, amateur radio is simply a vast body of knowledge and experimentation, and I find that I am constantly thirsting for new insight, projects, and information to fill in the gaps in my working knowledge and enhance the pleasure and enjoyment of our great hobby.

Over the years I have devoted a significant part of my time into exploring, investigating, and experimenting with the theory and practice of ham radio, and I am honored to be able to speak at the local clubs, the big Pacificon convention, as well as gatherings like the annual Digital Communications Conference. In speaking to audiences of several dozen to several hundred people, I have been encouraged, through suggestions of people in the audience, to publish my thoughts, findings and projects in a simple and down-to-earth newsletter format that with a mouse click I can send to thousands of interested hams at quarterly intervals.

What you will find in these pages, as the issues go by,

Using the internet to share and communicate information

are ideas and discussions over a range of topics that come from my own analysis and experiments. There will be no political or off-topic issues discussed, and nothing will be published that is a matter of controversy. There are plenty of forums for that, and I ultimately find them to be tedious and contentious. The topics you will find here are the same practical ideas and handy information that got you excited about amateur radio in the first place—theory, antennas, projects, kits, electronics, new technologies and much more.

We welcome your comments and suggestions.

Inside this issue:

Points of Interest:

- Our email address:

hampost@softwarescience.net

- Letters to the Editor— page 20
- Downloads — page 20
- Copyright policy—page 20
- Advertising policy—page 20
- Article submission policy —page 20

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Setting up a PSK31 Station

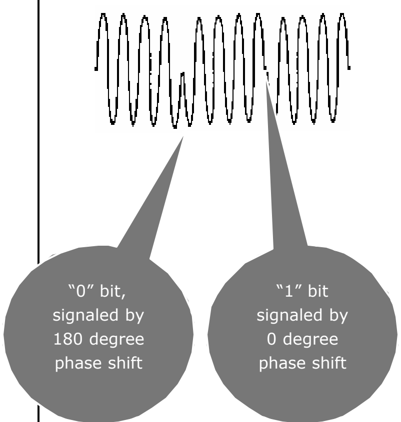


Figure 1. Basic principle of Phase Shift Keying

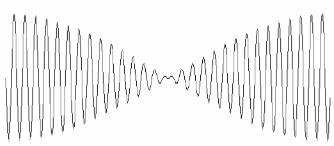


Figure 2. Phase reversals occur only during signal nulls

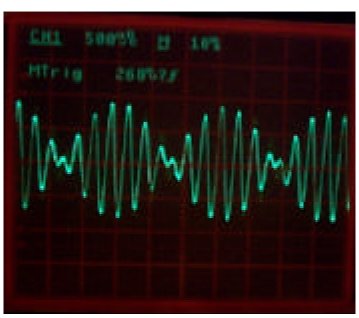


Figure 3. Scope photo of an "idling" PSK31 signal

PSK31 is one fun digital mode. To get into PSK31, you just need a computer with a sound card, an audio cable and your HF receiver. The fun of tuning with a waterfall display (instead of a tuning knob) and "listening" to QSOs in progress is well worth the few minutes to setup a PSK31 station. But first, let's go over some theory (not that you have to, but it sure is nice to know what is happening in your station).

PSK31 is a digital mode that transmits and receives bits based on shifting the f carrier's phase 180 degrees. Bits arrive every 32 milliseconds, so at 32 msec intervals we check to see if the phase is the same (a "1"bit) or has been flipped 180 degrees (a "0" bit). This idea is illustrated in Figure 1. The problem are those nasty sudden reversals in phase—they are going to produce wide bandwidth splatter! But if we gradually take the signal amplitude down to zero when we want to reverse the phase, there will be no sudden signal changes, and nothing splatters outside of the desired bandwidth. See Figure 2. Figure 3 is a photo of an actually oscilloscope trace of an "idling" PSK31 signal, i.e., sending nothing but zeros.

Sending bits every 32 msecs means we are transmitting 31.25 bits per second (aha—the "31" in PSK31), which translates into about 50 words per minute, about the fastest you can type.

One really great thing about PSK31 is its narrow bandwidth. Imagine putting 100 watts into a SSB voice signal spread over 3KHz. Putting the same power into PSK31, however, packs 100 watts into just 31 Hz. That's almost 100 times brighter (20 dB) than SSB! Another way to put it: you need much less power to have the same S/N ratio as voice. Cross-country contacts with 1 watt of power are not uncommon! PSK31 is ideal for QRP.

You are probably used to the ASCII standard, which encodes every character as 8 bits, with a start and stop bit.But PSK31 does not use ASCII! Instead, PSK31 uses a form of encoding whereby the most common characters are represented by the fewest bits, called "vericode". The average number of bits per character in vericode is about 6.5, compared to 10 for ASCII. This improves the data rate by about 50% for the same bandwidth! By the way, PSK31 allows both upper and lower case letters, numbers, and all punctuation (including backspace).

Let's Build One

Enough theory for now. Let's put a PSK31 station together. You probably have just about everything you need in your shack to assemble a monitoring station, so let's start with that.

1) First, you need a computer. I know you've got one—you might be reading this on it right now. Any computer with a Windows operating system will work (although PSK31 software is available for DOS and Linux also).

2) Next, you need a

"Cross-country contacts with 1 watt of power are not uncommon."

SoundBlaster compatible sound card. If you don't have one, you can get a low-end sound card easily enough for under $20.

3. Of course you will need an HF rig. Frequency stability (<1Hz) is a must. All modern solid state rigs meet this requirement. 20 meter capability is also important, since almost all PSK31 activity takes place on 20 meters.

4. You will need an audio cable with 1/8" mono or stereo connectors on both ends (e.g., Radio Shack 42-2497), that connects from your rig's speaker output to your sound card's line input.

5. And finally, you will need to install some PSK31 software. The best way to get started is with Skip Teller's (KH6TY) freeware DigiPan program. You can download a copy of Digipan from the Digipan Download Page (http://mywebpages.comcast.net/hteller/digipan/ and if you are interested in the technical details of Digipan, you can download Skip's June 2000 QST article (http://www.qsl.net/kh6ty/digipan.pdf) or Steve Ford's WB8IMY May 1999 article in QST "*PSK31-Has RTTY's Replacement Arrived?*".

Step by Step

Download, install and run Digipan. Tune your rig to 14.070MHz USB. Set the rig's audio level to mid range. Connect from the speaker output of the rig to the line in of the sound card (this works because the impedances of both devices are pretty close).

Open the Windows sound mixer by double clicking on the yellow speaker icon)—see Figure 6a. Click on Options|Properties, and then select "Recording"—see Figure 6b. Be sure to select Line-In and *deselect Microphone!*. Set the volume sliders for Recording and Line-In to mid-range. See Figure 6c.

"You probably have just about everything you need in your shack to assemble a monitoring station"

That's it! If you have done things correctly, you should see a waterfall display like the one in Figure 4.

The waterfall is a plot of signal intensity vs frequency, and covers about 3KHz of spectrum—your rig's audio bandpass. You will notice one or more

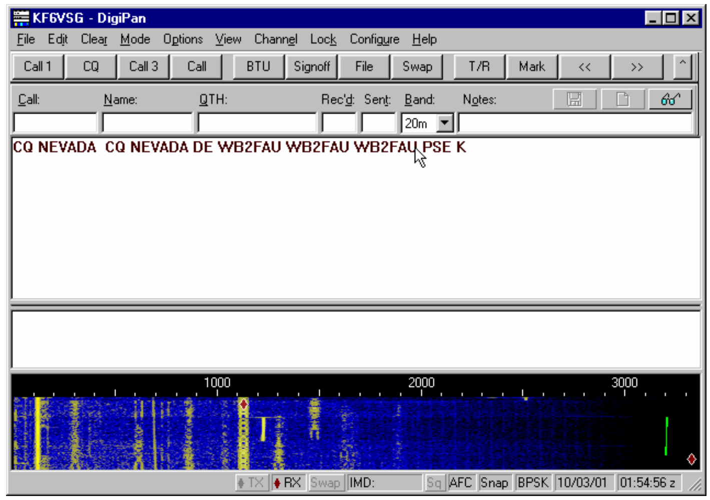


Figure 4. The Digipan program in action, receiving PSK31 signals. Notice the "tracks" in the waterfall display. Each track is a QSO in progress.

"tracks" or traces in yellow, each of which is a PSK31 signal. Adjust your rig's volume to give the best contrast between the yellow tracks and the blue background, usually when you just start to see sparkles in the waterfall.

Using your mouse, click on any of these tracks. Then watch the text area as the signal you choose is decoded into plain text. What you are monitoring is a QSO in progress! Of course, it won't be long before you will want to get on the air and enjoy transmitting, too.

On the Air

To transmit, you will need to route the line out of your sound card to the microphone input of your rig. This poses two problems that need a solution:

1. A sound card's line out is low impedance (it thinks you will be connecting to a 4Ù or 8Ù speaker), but your microphone input in high impedance. We need something to match impedances.

2. What is going to key up your transmitter?

Without going into all the details, the bottom line is that you will need to build or buy a "rig-to-computer" interface." Let me help you decide what to do in two ways. First, take a look at Figure 5. This shows a table of four the leading rig interfaces on the market. I have found that the very simplest interface (and least expensive), the NOMIC, from West Mountain Radio does a perfectly fine job to couple my sound card to my rig. On the other hand, the SignalLink interface from Tigertronics does not required an RS-232 interface or cable.

Secondly, Peter Halpin PE1MHO, Basil Helman G4TIC and Simon Brown HB9DRV have done an absolutely splendid job of explaining audio interfacing, complete with circuit diagrams and construction details, in their article "*Interfacing: A basic guide to CAT and audio interfacing*", available at http://sysgem.decus.ch/hb9drv/Interfaces.pdf. If you enjoy constructing your own interface (of out odds-and-ends found in your junk box), this is the article for you (just skip past the first sections on CAT interface and get right to audio interfacing).

So, assuming you have built or acquired an interface, the rest is easy:

1) Connect an audio cable from the line out (or speaker out) of your sound card to the interface audio input.

2) Open the Windows sound mixer (double click on the yellow speaker icon)—see Figure 6. Deselect everything but Wave. Set the Wave and Volume sliders to mid-range.

2) Connect an RS-232 cable (Radio Shack 26-117) from an unused serial

Product	Manufacturer	MSRP	Web site URL	Front view
RigBlaster	West Mountain Radio	$90	www.westmountainradio.com	
Nomic	West Mountain Radio	$35	www.westmountainradio.com	
SignaLink	Tigertronics	$50	www.tigertronics.com	
MFJ-1275 Soundcard Radio Interface	MFJ Enterprises	$89	www.mfjenterprises.com	

Figure 5. Table of leading rig-to-computer interfaces.

COM port on your computer to the DB-9 connector on the interface (if you purchased the Signal Link interface, you don't need this cable).

3) Connect a cable from the interface output to the microphone input of your rig. Now this is where things get tricky. Your rig has a special jack for the microphone, with an additional pin for PTT. Your interface has its own jack. Hence, a special cable has to be fabricated that not only has the right connectors on each end, but also the proper wiring to assure that audio gets to the mic pin and switch closure gets to the PTT pin. When you buy an interface, be sure to tell the manufacturer what rig you plan to use the interface with—they will then send you the appropriate cable ready to go. If you are building your own interface, you are on your own (although the article by Halpin *et. al.* gives details for some of the more popular rigs).

One of the most frequent complaints I hear is the need for yet another COM port. There are two ways around this. The first, as mentioned above, is to go with the SignalLink interface, which uses VOX to control PTT, and not a serial port. The second is to purchase a serial-to-USB converter, which lets you use a USB connection instead of a COM port. This ingenious device will be discussed in the next issue of *HamPost*.

In Digipan, set the COM port parameter in Configure|Sound Card to match the COM port you are using (if you are using SignalLink, set this to NONE).

That should be all you need to try out your first CQ. Just click on the "T/R" button in Digipan. Your rig should start transmitting the PSK31 idle signal. Now start typing CQ CQ CQ DE <your call sign> several times and click on "T/R" to return to Receive mode. If you have done everything correctly, you will probably

"One of the most frequent complaints I hear is the need for yet another COM port."

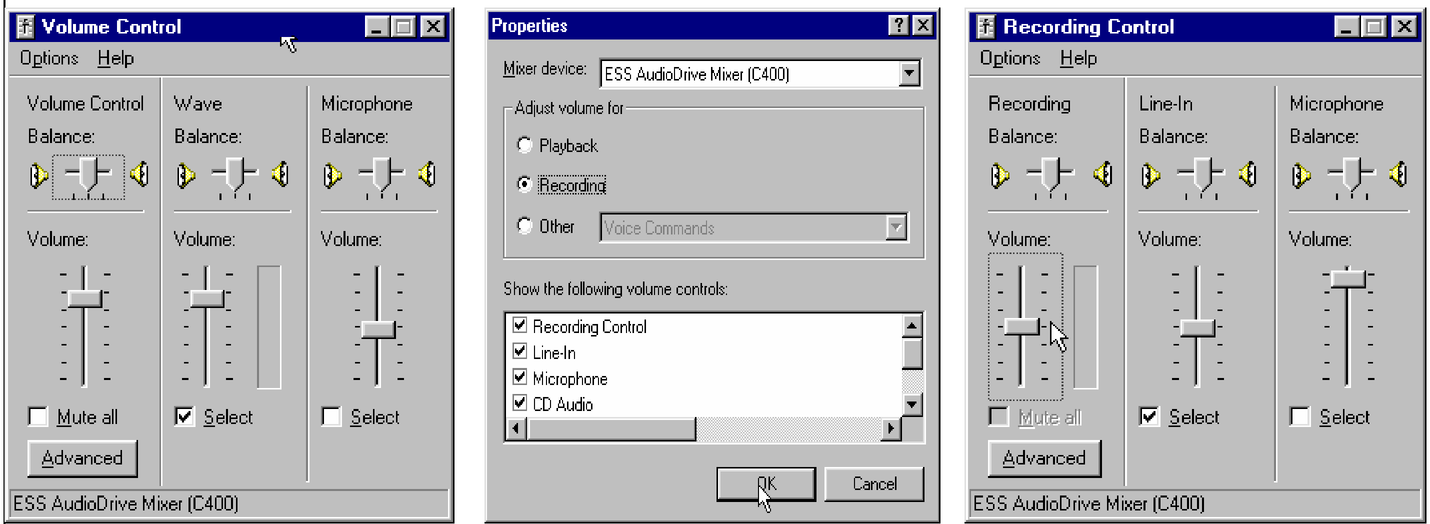


Figure 6. a) The Windows mixer. b) Select "Recording." c) Be sure to have only Line-In selected.

get a response—just look for the come-back in the text box.

Of course you can work it the other way—click on tracks in the waterfall until you see a CQ coming in. When the caller is finished, the track will disappear, and it is your turn to click on "T/R" and respond as you would to any CQ.

I assure you that the first time you make a contact in PSK31, the thrill will definitely be perceptible, perhaps approaching joy.

You are now a PSK31 (or PSK63) operator, and have hours of pleasure ahead of you. The next article in this issue of *HamPost* delves into an important and common issue—proper sound card level settings to insure that your signal is undistorted.

> ***"…the first time you make a contact in PSK31, the thrill will definitely be perceptible…"***

In the next issue of *HamPost*, I will discuss a number of QSO conventions peculiar to PSK31, such as abbreviations and macros, case, error correction issues, RST and IMD reporting, and some of the most common failure modes, and how to debug them.

For more information...

For more information on PSK31, try some of these technical articles from the pages of QST and QEX.

Issue	Page	Title	Author
Apr 2001 QST	83	Revisiting "Clean Up Your PSK31 Signal	Kruis, Richard J., K8CAV
Mar 2001 QST	37	The Warbler--A Simple PSK31 Transceiver for 80 Meters	Heron, George, N2APB
Feb 2001 QST	88	Clean Up Your PSK31 Signals	Kruis, Richard J., K8CAV
Oct 2000 QST	81	PSK31 on FM (Technical Correspondence)	Mason, Butch, W6KAG
Jun 2000 QST	31	A Panoramic Transceiving System for PSK31	Teller, Howard, KH6TY
May 2000 QST	45	"My PSK31 Doesn't Work!" (sidebar to PSK31 2000)	Ford, Steve, WB8IMY
May 2000 QST	42	PSK31 2000	Ford, Steve, WB8IMY
Dec 1999 QST	35	A PSK31 Tuning Aid	Urbytes, Don, W8LGV
Nov 1999 QEX	59	PSK31: A New Radio-Teletype Mode (Jul/Aug 1999 QEX)	Martinez, Peter, G3PLX
Jul 1999 QEX	9	Is PSK31 Legal?	Sumner, Dave, K1ZZ
Jul 1999 QEX	3	PSK31: A New Radio-Teletype	Martinez, Peter, G3PLX
Jul 1999 QEX	2	PSK31 (Empirically Speaking)	Smith, Doug, KF6DX
May 1999 QST	41	PSK31 - Has RTTY's Replacement Arrived	Ford, Steve, WB8IMY

Websites:

http://www.psk31.com

http://www.aintel.bi.ehu.es/psk31.html

http://www.arrl.org/tis/

Optimizing your PSK31 Signal

In a previous article we saw how to assemble and configure a PSK31 station, and hopefully you started enjoying contacts and filling up your log book.

No doubt you discovered that you could optimize the waterfall display for greatest contrast and best text readability by adjusting your rig's AF volume control, or your sound card's Line-In slider on the Windows mixer.

In the same way, you might have experimented with the Wave slider in the mixer in an attempt to boost your signal, transmitting as much power as possible. If you did, you probably got back less than complimentary signal reports, and requests to decrease your audio level. What's happening here?

PSK31 is very sensitive to the audio level setting. It is all too easy to over-modulate your rig, creating harmonic distortion and lots of splatter.

If you have a power meter, here's a simple experiment that will help to understand the problem. Feeding into a dummy load, measure the power output as you increase the Wave audio level from 0% to 100% of maximum audio. Figure 1 shows the results of my experiment, measured at 1500 Hz. Notice that the output power does not seem to increase for audio levels greater than 75% of maximum. This is a clear indication that we are getting out of the range of a linear response, and bad things are happening. Figure 2 is a scope trace of the RF output in this non-linear region. Just look at that choppy, flat-topped signal—imagine the huge splatter this would create, not to mention that the readability of the copy will suffer.

"…you probably got back less than complimentary signal reports… What's happening here?"

Figure 3 show the same RF output, where the audio level has been lowered to be well within the linear range—a perfectly modulated signal with no visible harmonic distortion. A signal like that will get you wonderful signal reports, and your copy will be as clear and readable as can be.

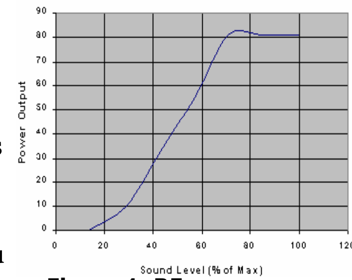


Figure 1. RF power vs. Audio level.

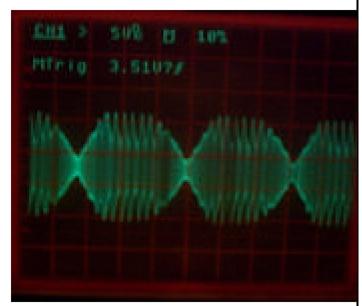


Figure 2. Oscillosope trace of over-driven PSK31 RF output.

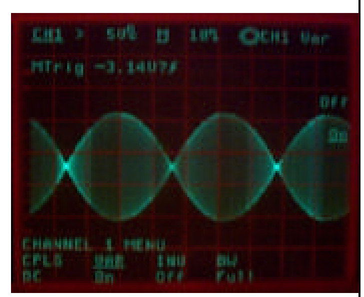


Figure 3. Having set the audio level to within the linear response range of the PSK31 system, this signal is clean and splatter-free.

Unfortunately you cannot tune up once and then forget the settings. The proper audio level depends on the audio frequency, that is, where you are in the PSK31 band. When you change frequencies, you should tune up all over again. Also, if others have access to your computer, you can never be sure the sound card settings have not been tampered with. Even some software applications can take control of the audio levels, leaving them in a state that forces you to re-tune.

One solution is to devise a circuit that will sample the RF output, compute the harmonic distortion, determine if the audio is set too high (or too low), and automatically adjust the audio level to assure a strong yet distortion-free signal.

Figure 4 shows a block diagram of a circuit that acts as the RF envelope sensor, providing a data stream to your computer via a serial port. The idea is to use the PIC 16F876 microprocessor, having an on-board ADC and UART. The front end of the ADC taps into your 50Ù transmission line, and the CPU samples exactly two PSK31 cycles, reporting 64 8-bit measurements to the computer through the UART interface. Shown in the figure are analog and digital scope traces through the signal processing steps. The figure's six captions contain further detail.

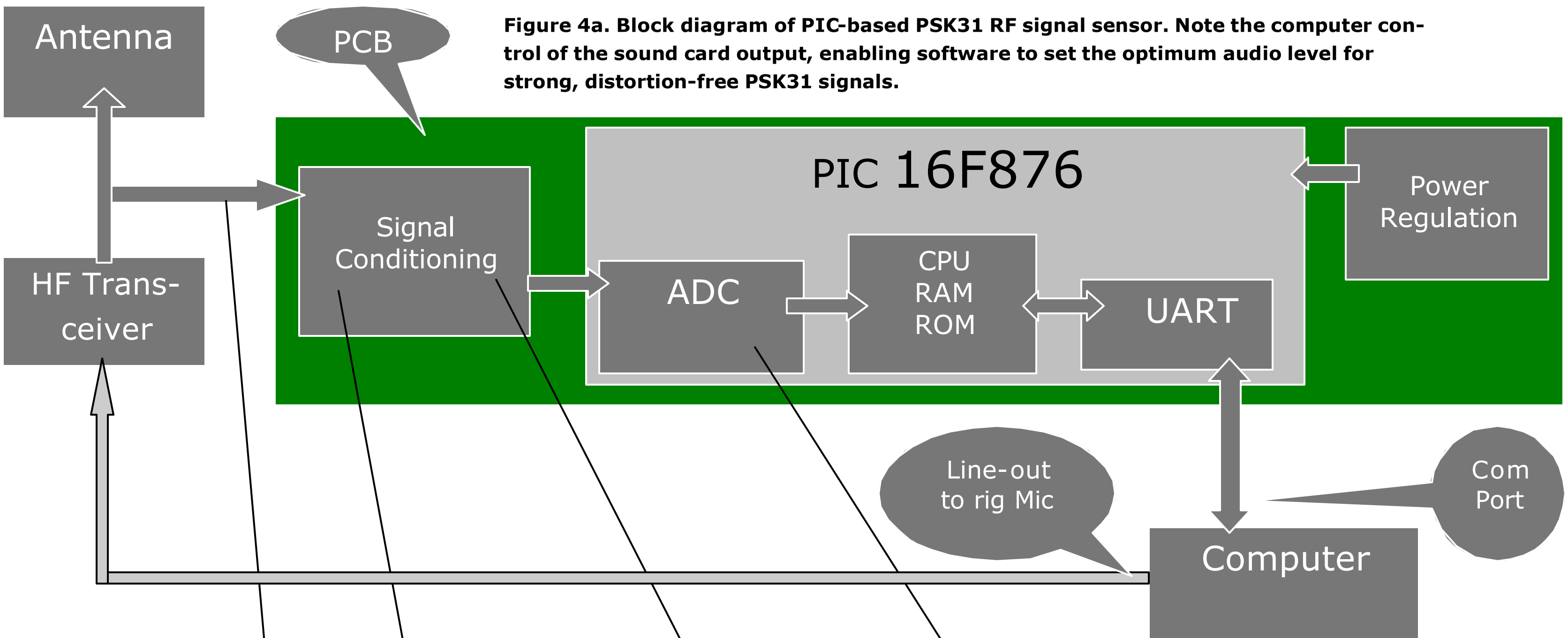


Figure 4a. Block diagram of PIC-based PSK31 RF signal sensor. Note the computer control of the sound card output, enabling software to set the optimum audio level for strong, distortion-free PSK31 signals.

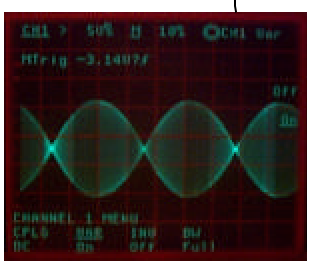


Figure 4b. PSK31 modulated RF signal as tapped from the feedline.

Figure 4c. RF signal after 1/2 wave rectification.

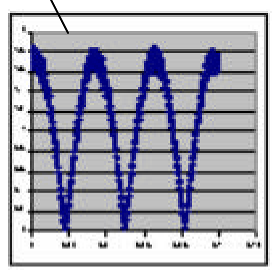


Figure 4d. Signal after low pass filter, ready to be digitally sampled

Time	ADC
0	236
1	248
2	236
3	248
4	231
5	236
6	219
7	214
8	191
9	191
10	163
11	163
12	129
13	118
14	84
15	78

Figure 4e. 16 ADC samples (1/2 of one PSK31 cycle). Time in milliseconds.

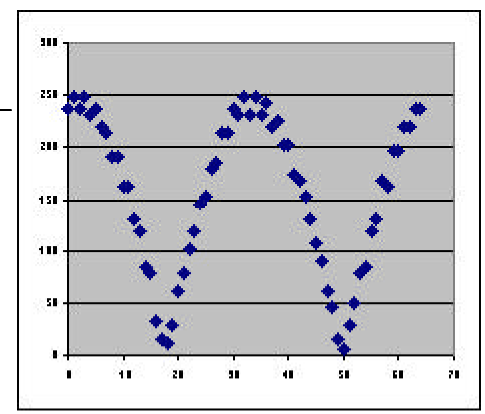


Figure 4f. Plot of ADC samples over 2 PSK31 cycles. This data forms a "packet" sent to a serial port of the computer for analysis.

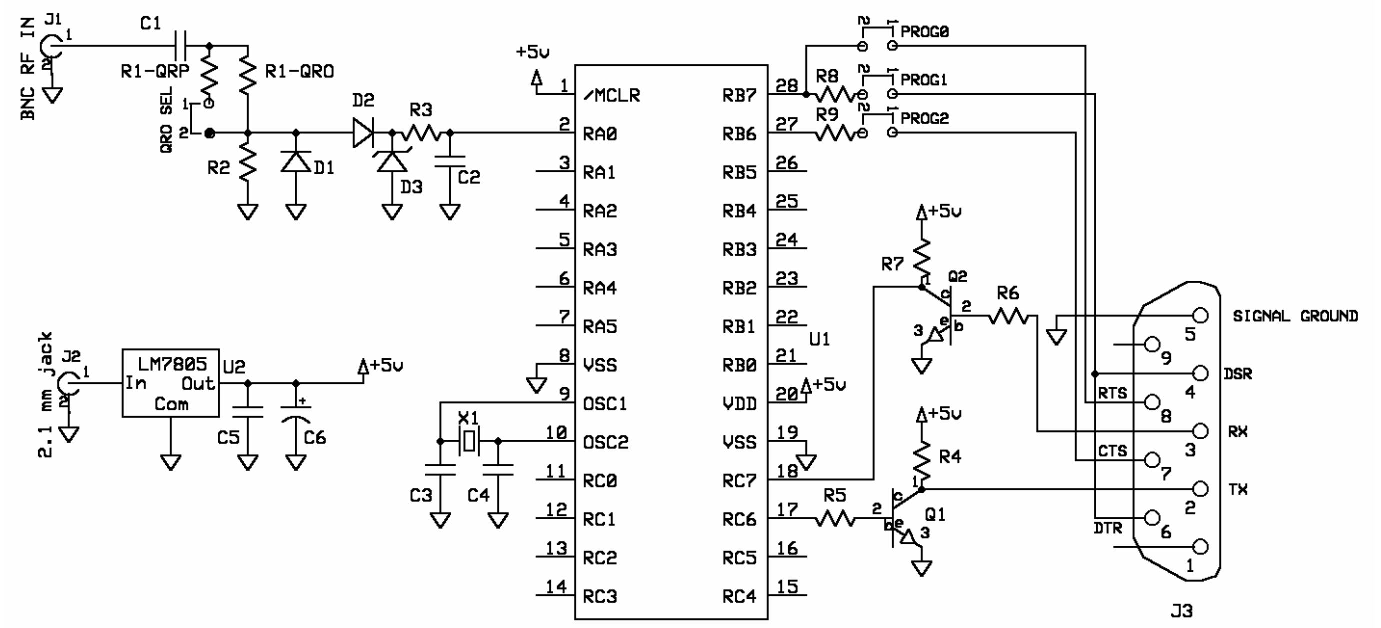


Figure 5. PSKMeter Circuit Diagram.

Although the signal processing illustrated in Figure 4 may seem complex, the actual circuit diagram and parts count for the sensor board is rather small. Figure 5 shows the complete circuit diagram for what has been coined the *PSKMeter*. The major circuit sections are:

1) RF input voltage divider, diode rectification and low pass filter,

2) PIC microcontroller with 20MHz crystal,

3) TX and RX level-shifting transistors (TTL to RS-232),

4) Power supply voltage regulation, and

5) Optional in-line programming signal, reserved for future use.

See Figure 6 for the parts list.

When constructing this circuit, be advised: you have RF signals and digital signals on the same board. That means that strong RF signals at 14 MHz (20 meters) will mix with the 20MHz crystal and shift the CPU's clock, causing the PIC firmware to crash.

The solution is to be careful with the physical layout of the board:

1) Keep all RF traces as short as possible,

2) Use plenty of ground plane on the copper side of the board, and even the component side also, and

3) Keep the PIC and its crystal away from the RF signal conditioning components and traces.

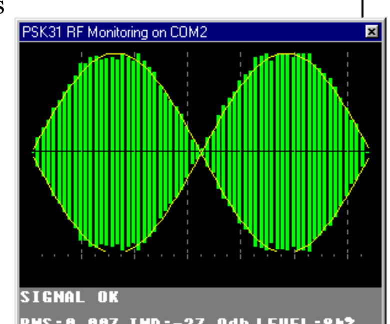
Layout can be tricky, but with perseverance I was able to obtain reliable performance. Figure 7 shows the parts layout of the final version of PSKMeter. The figure does not show the placement of ground, but I used as much ground plane on both the component and solder sides of the board as possible.

Figure 8 is a photo of the final assembly of the board.

The firmware for the PIC responds to commands received on the serial port. When instructed to do so, the PIC buffers 64 samples the audio modulation of the RF signal, spaced 1 millisecond apart. Thus, 2 complete PSK31 "bits" are captured and then transmitted to the computer. The PIC then waits for another command.

The software application on the Windows computer then receives this buffer and processes the data like this:

- normalizes the waveform to unity amplitude
- Phase shifts the waveform so that the zero crossover occurs in the first channel
- Measures the rms deviation from an ideal sine wave
- Measures the fundamental and first 10 harmonics, computing the IMD (intermodulation distortion)
- Presents the data graphically and displays the figures of merit :



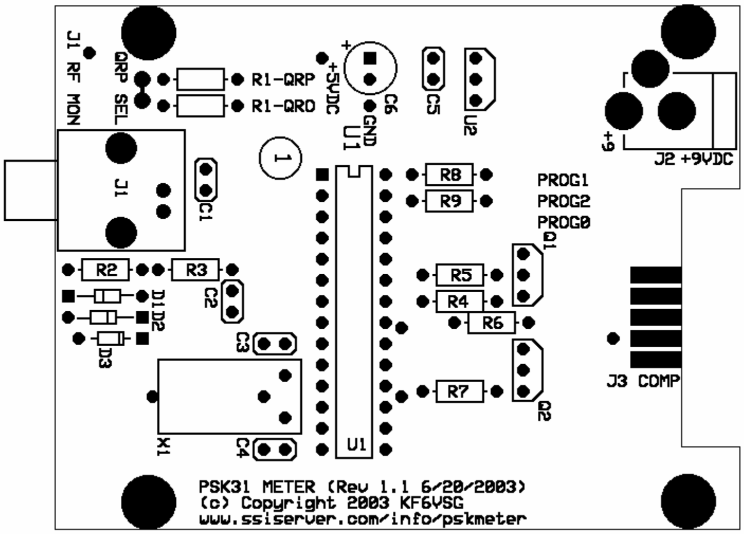


Figure 7. PSKMeter Layout

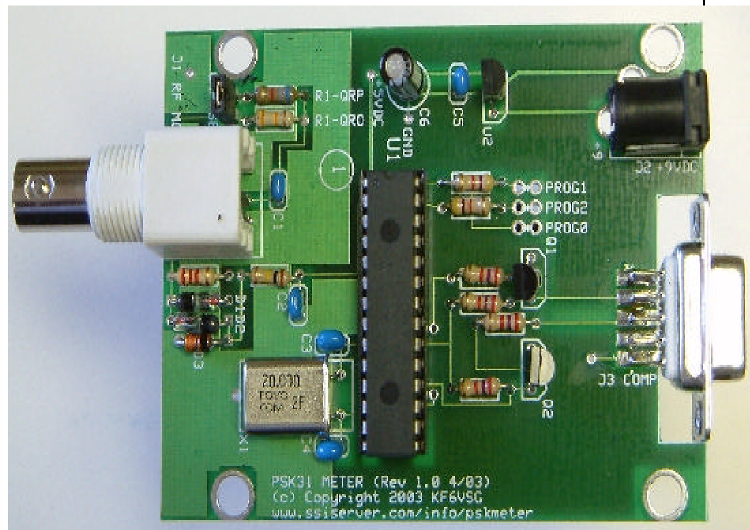


Figure 8. An assembled PSKMeter

R1-QRO	39K
R2-QRP	6.8K
R2	2.2K
R3	100K
R4,R5,R6,R7	2.7K
D1,D2	1N34A (or NTE 109) Ge diode
D3	5.1V Zener
X1	20MHz xtal
C2,C3,C4	22 pf monolithic
C1,C5	0.1 uf monolithic
Q1,Q2	2N4401 NPN
U2	LM7805
J3	DB9 Female edge connector
C6	10 uf electrolytic
J2	2.1 mm power
J1	BNC connector
R8,R9	4.7K
U1	PIC 16F876

Figure 6. PSKMeter Parts List

Once the application has computed the transmitted signals IMD and other figures of merit, an estimate is made of change in audio level required to improve either the signal strength (by increasing the audio) or the signal IMD (by decreasing the audio). By periodically (once per second) analyzing the transmitted RF and controlling the sound card's audio output to your transmitter, both weak signal and splatter are eliminated. The PSK31 station signal is therefore always optimized in the background.

"PSKMeter automatically finds the "sweet spot" in the audio level setting that outputs the highest power from your transmitter without creating splatter."

Additionally, the pskmeter.exe application lets you configure and control the following properties:

- Com port selection
- Mode selection (PSK31/PSK63)
- Refresh (sampling) rate
- Force window to be always visible
- View data numerically
- Force automatic control of audio level
- Feedback Gain selection
- Desired (target) IMD selection
- Noise (floor) threshold
- Maximum power emission selection
- Sound card selection

Building one for yourself

Using the parts list and circuit diagram, you can build your own PSKMeter. The binary (hex) file for the PIC, and the Windows program pskmeter.exe are freely available at no charge at `http://www.ssiserver.com/info/pskmeter`.

We have also made PSKMeter available as a complete kit including printed circuit board (with silkscreening and soldermask) at the same website as above.

For more information, see

QST, February 2004, *Short Takes,* page 66, Steve Ford WB8IMY

Radio Basics—Resonance

After Ohm's Law, nothing is more basic and pervasive in understanding AC circuits and radio communication than the principle of electrical resonance.

An understanding of resonance can be applied to a huge list of subjects, from oscillators, tuning circuits, and transmissions lines, to antennas, loading coils, traps, impedance matching and baluns.

In this issue of the HamPost, we discuss the basic principles, presenting a notation that can be appied over and over again to various situations, with the goal of providing the reader with the ability to apply these guiding principles to everyday practical situations. In future issues of *HamPost* we will build on these principles hoping to cover many topics in radio science.

Getting started

The place to start is with the humble capacitor and inductor in series. As you will soon see, there is a lot of magic that comes from this simple circuit.

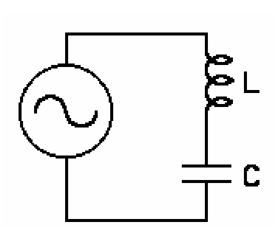


Figure 1. Series LC

The voltage and current in the circuit are related by Ohm's Law:

$$V = I \times Z$$

So that if we know the impedance, we know everything there is to know about this circuit. The impedance seen by the AC source, looking into this simple LC network is just the series sum of the impedances of the two components:

$Z = Z_C + Z_L$.

But $Z_C = 1/j\omega C$ and $Z_L = j\omega L$,

where j is the imaginary number (see sidebar), ω is the angular frequency in radians/sec (just 2ð times the usual frequency in Hz), C is the capacitance and L is the inductance. Because $1/\omega C$ has units of ohms, we say that $1/\omega C$ is the capacitive reactance (X_C) and since ωL has units of ohms, we say that ωL is the inductive reactance (X_L).

So that putting this together gives

$$Z = j(wL - 1/wC) \qquad (1)$$

which you might have seen many times. Of course, having seen it is certainly not the same as knowing what it means. So let's investigate some of the amazing implications of this result.

The Resonant Frequency

First, notice that because of that little minus sign it is possible for the impedance of the circuit to be zero, i.e., the circuit has no resistance at all to the flow of current! This happens at just one particular frequency, when

$$\omega L = 1/\omega C$$

(when the inductive reactance equals the capacitive reactance). Solving for ω gives

$$w_0 = \pm \frac{1}{\sqrt{LC}} \qquad (2)$$

Notice that we give this special frequency a special symbol, ω_o, and a special name: *the resonant frequency.* Usually we only consider the positive solution for ω_o, but as you can see there is also negative ω_o. For instance, if the resonant frequency is 1000 Hz, the circuit will also resonant at –1000 Hz. Of course this makes us wonder, what is –1000 Hz? This is just the same signal as a 1000 Hz signal but phase shifted 180 degrees. A way to think about this is that if a circuit is resonant at a certain frequency, it will also be resonant if you put the input of the circuit through an inverter.

"There is a lot of magic that comes from this simple circuit."

Now something very bizarre happens when we apply AC at this frequency to the circuit: the impedance utterly vanishes—the circuit becomes a dead short. In fact, the circuit just starts soaking up all the energy you can throw at it. You know from physical objects in real life that if you shake an object

That wacky number, j.

Since j is $\sqrt{-1}$, it has these unusual properties:

j has no dimensions, it's a pure number

$$j^2 = -1$$

$$j \times (-j) = 1$$

$$1/j = -j$$

This is all the math you have know about j. You can use these simple facts whenever you solve equations that involve imaginary numbers.

at just the right frequency, that object is going to stop resisting you, and start oscillating with larger and larger amplitudes. When that happens, we say the system is resonating, or exhibiting resonance, and we have hit the resonant frequency.

From here textbooks discuss the impedance away from resonance, and phase relationships and a lot of complicated looking formulae. I want to you to see that it is possible to make the whole thing a lot simpler.

Simplify, Simplify...

One thing we can do right from the start is to find ways to "normalize" equations and formulas. When a formula is normalized, we express things as ratios to some standard quantity, and a lot of the clutter vanishes, and on top if it we get equations with lots of dimensionless quantities (just pure numbers) which are much easier to work with and think about. Let's try it. Equation (1) can be rewritten like this:

$$Z = j X (\mathbf{w} - 1/\mathbf{w}) \qquad (2)$$

Where the X is the reactance of the inductor (or capacitor) at resonance ($=L\omega_0$), and the bold $\mathbf{w}$ is actually the ratio of ω to the resonance frequency ω_0. See the side bar for details.

ω_0 is a very special quantity, because we can use this property of vanishing impedance for lots and lots of applications. Therefore, it makes a lot of sense to talk about the bold frequency $\mathbf{w}$ as the ratio of frequency to the resonant frequency. It means that this bold frequency has <u>no units</u>, not even Hz or cycles, or radians/sec, or anything. It's just a pure, dimensionless, number. But we can always go back to Hz just be multiplying it by ω_0. For example, whenever we see that $\mathbf{w}$ is 1, it means we are talking about the resonant frequency. Unlike the unnormalized quantities, the normalized resonant frequency is always 1, for any circuit, any value of L or C, and so on. Similarly, if we say $\mathbf{w} = 2$, that would be the first harmonic of the resonant frequency, and so forth. So that is what happened when we normalized the frequency.

Also notice that X appears, the reactance of the inductor at resonance ($\omega_0 L$). This might seem odd: we started with an equation (1) that was kind of symmetrical with L and C, but in equation (3) the value of the capacitor disappeared! However, since at resonance the reactance of the inductor is exactly equal to the reactance of the capacitor, would could have just as well written equation (3) this way:

How'd he do that?

Start with equation (1) and factor out $L\omega_0$:

$$Z = jLw_0\left(\frac{w}{w_0} - \frac{1}{LCw_0 w}\right)$$

Remember that $LC = 1/\omega_0^2$ (from equation 2). So substitute that in and you'll get:

$$Z = jLw_0\left(\frac{w}{w_0} - \frac{1}{w/w_0}\right)$$

Let $\mathbf{w}$ stand for frequency normalized to the resonant frequency, i.e., $\mathbf{w} = \omega/\omega_0$ and call $L\omega_0 = X$, and voila, you have equation (3).

$$Z = j X_C (\mathbf{w} - 1/\mathbf{w})$$

And now the inductor disappears! Because $X_C = X_L$ at resonance, it doesn't matter a bit which we use, so we use the symbol X to represent either one.

Another name for this special value of X is the "characteristic impedance" of the circuit. Probably you have heard this term used for transmission lines (like coax). It's called characteristic, because it is the only value in ohms that drops out of these equations all by itself. It looks like X is dependent on frequency, but actually if you substitute the value of ω_0, you will get

$$X = \sqrt{\frac{L}{C}} \qquad (4)$$

So the characteristic impedance just depends on the electrical properties of the circuit and is independent of the signal you put into it. Later on I will have more to say about characteristic impedance and some special magic associated with it.

For now let's make one more step, and normalize the impedance Z, by dividing both sides by X to get

$$\mathbf{Z} = j (\mathbf{w} - 1/\mathbf{w}) \qquad (5)$$

But now you see that **Z** is bold, meaning it is the impedance normalized to the characteristic impedance. That is, **Z** is now also dimensionless, and is a pure number, and is not in ohms or mhos or anything like that. To get back to a physical value, we just have to multiply **Z** by the characteristic impedance, X.

So equation (5) is the impedance equation for every series LC circuit that ever was soldered together and because it is normalized, it is reduced to the simplest way of expressing the relationship of the variables. Amazingly we have gotten rid of L, and C, and there is not a Hz, second, ohm, henry or farad in sight. So, doesn't it look so much more clear and crisp than equation (1) or (2) where we had to plug in real-world values of L and C? Of course equation (5) isn't that wildly different, but wait until we get to more complex circuits—the difference will be really convenient—and amazing.

Just by looking at equation (5) we can immediately grasp lots of physical conclusions that have to be true for all LC series circuits:

> Except at resonance, the impedance is always complex. **Z** will always be either a positive number times j, in which case the circuit is equivalent to nothing more than an inductor, or a negative number times j, in which case the circuit is equivalent to nothing more than a capacitor. That is, away from resonance, you can always find an equivilent inductor or capacitor.
>
> **Z** =0 when $\mathbf{w} = 1$. At resonance the current will tend to infinity, and it will be exactly in phase with the voltage.
>
> When $\mathbf{w} < 1$, the circuit has negative impedance, i.e., its inductive. The voltage leads the current by exactly 90 degrees.
>
> When $\mathbf{w} > 1$, the circuit has positive impedance, i.e., its capacitive. The current leads the voltage by exactly 90 degrees.
>
> At DC, $\mathbf{w} = 0$, so $1/\mathbf{w}$ is infinite, and **Z** will become infinite. No current will flow.
>
> At infinitely high frequencies, **Z** also becomes infinite. No current will flow there, either.

"...we didn't have to even think about specific capacitors and inductors, or units, or compute numbers or anything messy like that."

Equation (5) let's us think about negative frequencies, too. What's a negative frequency? Well, it's the same physical oscillation as the positive frequency, but 180 degrees out of phase. You can see from equation (5) then that when $\mathbf{w} = -1$, we *also* get **Z**=0, that is, there is another resonance. And that's common sense: the circuit will resonant at the resonant frequency and if I invert the signal so its shifted 180 degrees, its going to resonant just the same.

What about when **Z** = j (or –j)? What would be the frequency be for the impedance of the circuit to equal the characteristic impedance? Just set **Z**=j and solve for $\mathbf{w}$:

$$j = j(\mathbf{w} - 1/\mathbf{w})$$

$$\mathbf{w} = \mathbf{w}^2 - 1$$

$$\mathbf{w}^2 - \mathbf{w} - 1 = 0$$

Using the quadratic formula, we can solve for $\mathbf{w}$:

$$\mathbf{w} = (1 \pm \sqrt{5})/2$$

$$= 1.62 \text{ and } -0.62.$$

What this means is that there are two frequencies that this happens at, one 62% above the resonant frequency, and one below the resonant frequency and inverted (or phase shifted 180 degrees). By the way, if you are mathematical, you will recognize 1.62 as Fibonacci's number (but how the 17th century Italian mathematician got into this discussion is beyond me).

So we see that we have a wealth of information pouring out of Equation (5), and we didn't have to even think about specific capacitors and inductors, or units, or compute numbers or anything messy like that.

Getting real

Can we make this circuit more interesting? We sure can. We can throw in some old fashioned resistance and see what happens. But notice there are four different ways to do that: we can just add some resistance in series, or we can put the resistance in parallel with the inductor, or in parallel with the capacitor, or in parallel with the whole thing. That means we can obtain 4 different formulae for the circuit impedance. In the next issue of *HamPost* I'm going to show you how easy it is to bat out the equations for all of these variations, but for now, let's just focus on adding a resistor in series:

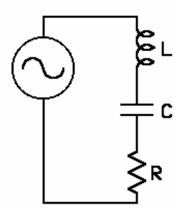


Figure 2. LCR circuit

$$\mathbf{Z} = j(\mathbf{w} - 1/\mathbf{w}) + \mathbf{R} \qquad (6)$$

Notice that we are now thinking in terms of normal quantities. That means when I add a resistor, its resistance will *not* be in ohms, but will be normalized to the characteristic impedance we talked about earlier, and its resistance will be just a pure, dimensionless number, which is denoted as bold **R**.

It would be nice to make a graph of **Z** vs w, but I don't know how to handle the mix of real and imaginary parts. The way around this is to get a grip on the magnitude of **Z**, and plot that vs w. We will lose information about the phase, and phase can be fun to think about, but for now let's just focus on the magnitude of the impedance. Here is what you get for the magnitude of **Z**:

$$|\mathbf{Z}| = \sqrt{\Omega^2 + R^2} \qquad (7)$$

where $\Omega = w - 1/w$, the reactance of the circuit. At resonance, $|\mathbf{Z}| = \mathbf{R}$ (because w is 1 and therefore Ω is zero).

We can now graph $|\mathbf{Z}|$, but what is more interesting is to graph the current in the circuit vs. frequency. From Ohm's Law, $I=V/|Z|$. Let the amplitude of the voltage be 1, so the current is just $1/|Z|$. So let's plot $1/|\mathbf{Z}|$. I used Excel, but you can crank out the numbers in Visual Basic, or whatever you feel comfortable with. My results for $\mathbf{R} = 0.05$ (i.e., the resistor is 5% of the characteristic impedance) came out like this:

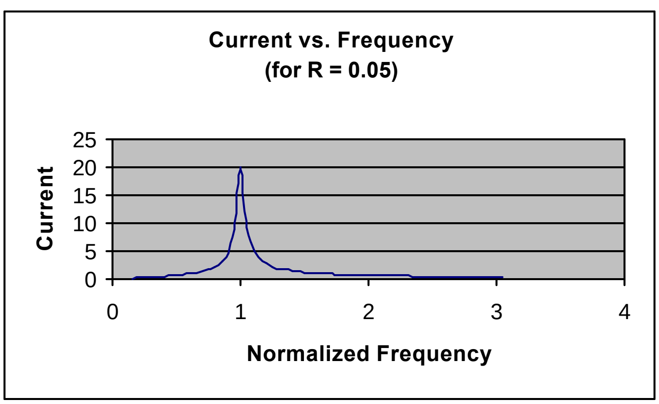


Figure 3. Current vs. Frequency

It will always be the case that you get a peak. The maximum will always be at $w = 1$, and falls off to zero almost symmetrically on each size of resonance. The peak value will always be $1/\mathbf{R}$ (can you see why?—go back to equation (7)). As **R** gets smaller, the peak gets narrower and higher, and as **R** gets larger, the peak flattens and gets shallower.

The next question to ask is what is the bandwidth of this circuit, that is, a measure of the width of this peak.

How'd he do that?

The square of the magnitude of any complex number is the product of the number times its complex conjugate. *Its what?* To find the complex conjugate, just change all the j's to –j's. That's all there is to it. The result of this multiplication is guaranteed to be a real, positive number. So:

$$|\mathbf{Z}|^2 = [jw + 1/jw + \mathbf{R}][-jw + 1/-jw + \mathbf{R}]$$

If you sit down and multiply all the terms out, you will find a lot of cancellation, and you will end up with equation (7). Try it.

Bandwidth (BW) is defined as the difference in the two points where the current is $1/\sqrt{2} = 0.707$ of the maximum current. Since the maximum current is $1/\mathbf{R}$, to get the bandwidth we just need to find the frequencies that result in $|\mathbf{Z}| = \sqrt{2}\,\mathbf{R}$. A glance at equation (7) tells us that $\Omega = \mathbf{R}$ gives us this result (no real math needed to figure it out). What this means physically is that bandwidth is the range of frequencies from the frequency below resonance where the circuit reactance (Ω) matches the resistance (**R**) to the frequency above resonance where the circuit reactance matches the resistance. So it's a very simple condition when we look at it that way. So how do we solve for BW, knowing that $\Omega = \mathbf{R}$? Since $\Omega = w - 1/w$, it means we are trying to solve

$$w - 1/w = \mathbf{R} \qquad (8)$$

This turns out to be a quadratic equation with two solutions (w above and below resonance), and it gets surprisingly messy. Let's use a mathematical trick to make it easier. Think of w as a number close to resonance, i.e., close to 1. Let Δw be the deviation from resonance, so that

$$w = 1 + \Delta w$$

That means that equation (8) can be written as

$$1 + \Delta w - \frac{1}{1+\Delta w} = \mathbf{R}$$

But $\dfrac{1}{1+\Delta w}$ can be approximated by $1 - \Delta w$

Putting this in gives us

$$1 + \Delta w - (1 - \Delta w) = \mathbf{R}$$

and this leaves us with

$2\Delta\omega = R$.

Now $\Delta\omega$ is the deviation on one side from resonance, so $2\Delta\omega$ is the deviation from below resonance to above resonance, which is exactly what we mean by bandwidth, hence we get the very beautiful result:

$$\mathbf{BW} = \mathbf{R} \qquad (9)$$

Now don't be fooled. Remember that **R** is normalized (it's bold, so its dimensionless), and <u>not</u> the value of the resistance in ohms. In fact, BW is also normalized, since we have been solving normalized equations all this time. If we want to know the bandwidth in Hz, we have to "undo" the normalization. Remember that to get back to real frequencies we have to multiply by the resonance frequency (ω_0), so

$$BW = \omega_0\, \mathbf{BW} = \omega_0\, \mathbf{R}$$

and **R** is the resistance R divided by that characteristic impedance, X. So in the books you will see it this way:

$$BW = \omega_0 R/X = \omega_0 R/\omega_0 L = R/L$$

Where here we show BW and R as their physical values (not bolded, not normalized). Still, the normalized result is a whole lot easier to compute with. Example: suppose the characteristic impedance of a circuit is 1000 ohms, and the resistance in the circuit is 50 ohms. Then **R** is 50/1000 = 0.05, so **BW** is 5%. Well, usually that's about all we wanted to know. The peak width is plus or minus 2.5% around resonance. If you know the resonant frequency, say 1 MHz, the bandwidth is 5% of that, or 50 Khz. Too easy.

So what have we learned? The most important thing about equation (9) is that bandwidth is proportional to the series resistance in your circuit. With no resistance (just LC only, made out of superconductors), the bandwidth would be zero. Actually that doesn't sound like such a good thing, does it? What good would a filter be with zero bandwidth? A little resistance is good. It means the resonance has some spread to it. And, if you double the resistance, you will double the bandwidth, and so on.

Also, by returning to physical properties, we see that you can always compute the bandwidth of the circuit by dividing the resistance by the inductance. For example, suppose the resistance is 10 ohms and the inductance is 1 mH = 0.001 henry. Then BW = 10/0.001 = 10,000 radians/sec. Radian/sec? Yes, because we have been working with frequency in that units. To get to Hz, just divide by 2ð, giving us a bandwidth 1,592 Hz.

The final quantity we can compute for this circuit is its "quality", or Q. Physically speaking, a circuit's quality is how long the circuit will "ring" after we remove the power source. For instance, when you strike a tuning fork, it continues to vibrate at the resonant frequency for a long time. If the resonance frequency of the tuning fork is 440 Hz, and it rings for 2 seconds before it damps out, the Q of this device would be 880. On the other hand, the Q of an automobile suspension system is designed to be low, maybe 1 or 2. For this reason, when you go over a bump, the car will oscillate only or two cycles. In the same way, when a circuit is resonant, and you remove the signal generator that is exciting the circuit, current will continue to flow in the circuit for additional cycles. The number of cycles is Q, the quality.

Q is just the inverse of **BW**:

$$Q = 1/\mathbf{BW}$$

So

$$Q = 1/\mathbf{R}$$

Because Q has no dimensions, to revert back to physical values, we just have to replace **R** with R/X and we get things the way you see them in the books:

$$Q = X/R$$

If you would like a tool for computing resonance circuit values, I have written a calculator which is freely available at

http://www.softsci.com/hamradio/download/V1N1-resonance.zip

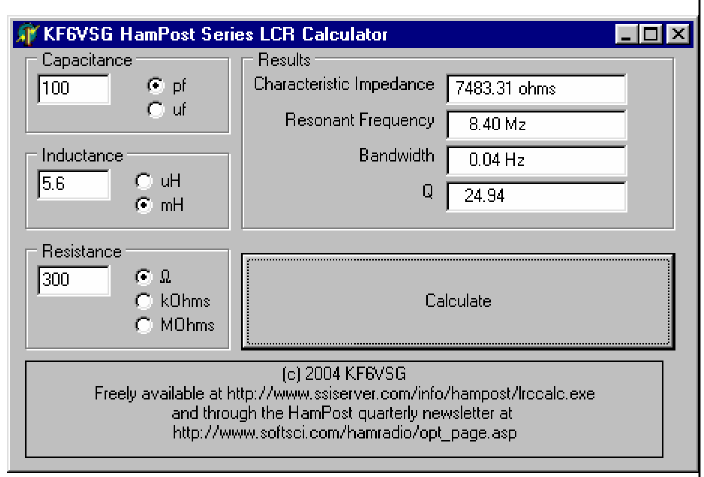


Figure 4. The LCRCalc Windows Program

Summing up

Let's summarize by writing down the basic formulae all in normalized units:

w = 1	(resonant frequency)
X = 1	(characteristic impedance)
Ù = **w** -1/**w**	(circuit reactance)
Z = j**Ù** + **R**	(complex impedance)
|**Z**| = $\sqrt{\Omega^2 + R^2}$	(magnitude of impedance)
BW = **R**	(bandwidth)
Q = 1/**R**	(quality)

Remember, every single quantity you see above is dimensionless. Just pure numbers. The beauty of this that each formula is simple and clean, and doing the math is easy. And keep in mind that these relationships apply to every series LCR circuit under the sun.

The association of normal quantities with the real world numbers is given by the rules:

1. To convert an impedance variable back to ohms, multiply by sqrt(L/C) .

2. To convert a frequency variable back to Hz, multiply by 1/2ð sqrt(LC).

3. Quality is dimensionless, so you don't have to do anything.

4. C (in farads) = 1/(ω_0X).

5. L (in henries) = X/ ω_0.

Example: An LCR AM Radio

Suppose we wanted to build an AM radio tuned to a fixed frequency using the circuit shown in Figure 5.

Recall that the impedance of the series LC is zero at resonance. So at the resonant frequency, the maximum current will flow through the resistor, and peak voltage appears across the resistor. The diode rectifies the RF AC, converting the signal to baseband audio.

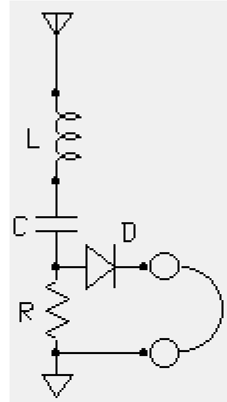


Figure 5. The LCR AM radio

Let the carrier frequency of interest be 1000 KHz (so $\omega_0 = 6.28 \times 10^6$), and we want to set the bandwidth to 50KHz (±25KHz). This means that Q=20 (1000/50), so **R**= **BW**=1/20=0.05. If we pick the resistor to be 300 ohms, that means that X must be 6000 ohms (because **R**=300/X=0.05). Using rules #4 and #5, we find that C=26.5 pf and L=955 µf. How easy was that?

Now run the lcrcalc.exe program that you downloaded and enter these values of L, C and R and verify you get the circuit to tune the desired frequency and bandwidth. Pretty cool, huh?

Another approach is to tinker with the values of L, C and R in lcrcalc.exe and see what you get for resonant frequency and bandwidth. There will be an infinite number of combinations of L and C that will yield the right frequency, but you will find only a limited set of values that also provide the desired bandwidth. In addition, you can only pick values of L and C that are manufactured and readily available.

Stay tuned...

In the next issue of *HamPost*, we'll build this classic circuit and even record the audio to the computer.

By now you are getting to be an expert in the series LCR circuit. To really get a grip on visualizing what is going on physically, we need to look inside this circuit and see where the energy is going. We'll pick up on this topic in the next issue of *HamPost* . Once we examine energy flow, its time to show how to transform this lowly circuit into a real-world antenna (yes—an antenna is just a distributed series LCR circuit!), using our results to learn about antenna resonance, bandwidth, Q, resonant length, characteristic impedance, trim factors, radiation resistance, antenna efficiency, with hints and tips for improving your antenna performance.

For more information

There is no end to the resources available on resonance, but here are few places to dig deeper:

ARRL, *ARRL Handbook* (any year)
Ian Poole, *Newnes Guide to Radio Communications Technology*
Raymond Serway, *Physics for Scientists & Engineers*

Understanding Path Loss and Signal to Noise Ratios

"You see, wire telegraph is a kind of a very, very long cat. You pull his tail in New York and his head is meowing in Los Angeles. Do you understand this? And radio operates exactly the same way: you send signals here, they receive them there. The only difference is that there is no cat. "
--Albert Einstein, when asked to describe radio.

Yes, there certainly is no cat. But there *is* an RF path, and that path plays a large role in determining whether or not the communication link will conduct information faithfully from the transmitter to the receiver.

A long way on a little geometry

What is the path loss? Imagine a transmitter with an isotropic antenna, so energy is flowing equally in all directions. Now suppose you place an antenna a distance r away from the transmitter. How much of the power that is radiated by the transmitter is captured by this antenna?

Suppose for now that your antenna is a parabolic dish, with a radius of ñ, pointed right at the transmitter. Its easy to see that the area of the dish, πñ², will capture all of the energy falling onto that area. But at a distance r, power from the transmitter has spread out and is crossing a big imaginary ball, with a surface area of 4πr². So the fraction of power captured by the antenna is just the ratio of the two areas:

$$P_r = P_T\left(\frac{pr^2}{4pr^2}\right) = P_T\left(\frac{r}{2r}\right)^2$$

Simple enough so far, but of course you probably don't use a dish antenna for your HF station, or HT or mobile. Instead, you use a "wire" antenna. But what then is the "aperture" of an antenna like that?

The answer is, you can still think of your dipole, for instance, as being the diameter of a disk, and waves that enter that disk are captured by your antenna. The "area" of a dipole would still be πñ², but ñ would be the length of one element of the dipole, which we know is 1/4 wavelength. Therefore, the

"The only difference is that there is no cat. "

first approximation to the aperature of a dipole would be πë²/16, where ë is the wavelength of the radiation we are receiving. It turns out that this is a bit of an overestimate, because the current that is flowing in the dipole has a cos(è) shape, where no current is flowing at all at the ends and lots of current flows in the center of the dipole (see the next issue of *HamPost* for insights on dipoles). If we average the value of cos (è) , we get an effective length of the dipole element of (ë/4)(2/π). So the aperture (area) of a dipole is about 40% of the disk you get by spinning the dipole about its center. By using this area we get

$$P_R = P_T\left(\frac{l}{4pr}\right)^2 = 6.33\times10^{-3}\left(\frac{l}{r}\right)^2 P_T$$

The term "geometric path loss" refers then to the dimensionless factor that links the transmitted power to the received power:

$$PL = 6.33\times10^{-3}\left(\frac{l}{r}\right)^2 \qquad (1)$$

Written this way, ë and r have to be in the same units (like meters).

In the books, you might see this written in terms of dB, which you get if you just take 10 times the log of PL:

PL(dB) = -22 + 20 log(ë) - 20 log(r)

In some texts, PL is expressed as a function of frequency in MHz rather than wavelength, and the distance in km. By substituting ë=c/f, and using km = 1000m, you get

PL(dB) = -32.45 –20 log(f) - 20log(r)

Which is the formula developed by John Pierce of Bell Telephone Laboratories (the same man who named the transistor), and is a result of pure geometry.

What can we immediately learn from eqn. (1) (or it's dB cousins)?

- When r = ë, the path loss is always -22 dB, regardless of the values of r and ë. That is, when you are one wavelength away from the transmit antenna, your receiver will pick up 0.63% of the transmit power at any frequency. This distance is also of interest because it is at the boundary between the near and far fields. Any closer to the transmit antenna, the inverse square law breaks down and our conclusions will be all wrong. Normally the distance is many wavelengths, so this is not of concern in real life.

- The power you receive follows the inverse square law: at twice the distance you will receive 1/4th the power.
- The power you receive goes as the square of the wavelength. Want to receive 4 times more power? Double the wavelength. Big antennas are good thing, after all!
- If you keep the ratio of wavelength to distance constant, the received power will not change.
- DX (large values of r) require proportionally larger values of wavelength (and thus lower frequencies). If you can receive a station from 1000 km on 10 meters, you should be able to receive a station operating with the same power and type antenna from 2000 km on 20 meters. This is the reason why DX is only possible on HF.

Voltage at the receiver

Next, let's consider what happens to this power as it enters your receiver. What is of interest is the voltage seen at the front end of the receiver. From Ohm's Law,

$$P_R = V_R^{\,2} / R$$

Where R is the input impedance of the receiver (or equivalently, the antenna impedance), and for radio amateur applications is almost always 50 ohms. You might wonder why I haven't including the power factor in the above equation. Recall that when the antenna resonates, the voltage and current are in phase, so we can use Ohm's Law as if we are working with DC. Solving for V_R and using eqn (1) gives

$$V_R = 0.56(l / r)\sqrt{P_T} \qquad (2)$$

where V is in voltage and P_T is in watts.

Example: Assume we are operating at 100 watts on 20 meters 2000 km from the receiver. What will the front end voltage be? Remembering to express r is meters, we get V=56 microvolts. That's useful to know, since we can look up the receiver's sensitivity (say its 5 microvolts) and immediately know that our signal to noise ratio is about 11:1. Useful information.

So far we have computed the power and voltage at the front end of the receiver. But will this mean we hear anything intelligible? To answer this, we have to compare the signal we are interested in to the noise that is competing to mask the signal.

Thermal Noise

Knowing the receiver's noise level will enable us to compute the Signal to Noise ratio (S/N) . Knowledge of S/N tells us whether our signal will be strong, marginal, or lost.

The equation for a solid state receiver's front end noise is

$$P_N = 4kTB \qquad (3)$$

Where P_N is the noise power, k is Boltzmann's constant (1.38×10^{-23} joules/Kelvin), T is the temperature in degrees Kelvin (293 at room temperature), and B is the receiver bandwidth in Hz. Because this so-called "thermal noise" is a theoretical limit, this is about the best (least) noise can be. A this point in time, no one has a devised a way to get rid of this noise, so we are stuck with it.

A quick look at eqn (3) tells us a lot:

- Amazingly, thermal noise doesn't depend on frequency. It's spectrum is flat as a pancake from DC to daylight. When a noise spectrum is flat, like that of white light (all colors present in equal quantities) we say it is "white noise."
- Noise power is proportional to bandwidth. You can reduce thermal noise by squeezing your signal into fewer Hertz, but the price is a lower information rate (we'll discuss bandwidth, channel capacity, and baud rate in a future *HamPost*).
- Noise power is proportional to absolute temperature. Unless you plan to keep your antenna and receiver front end at liquid nitrogen temperature, we set T=293 and consider it a constant.

You can hear thermal noise anytime you like, just by opening the squelch on your receiver, and disconnecting the antenna. Turn up the volume and you will hear a steady hiss. That's it--thermal noise, amplified and demodulated to baseband. Thermal noise is just a mix of random amplitude, random frequency

"...no one has a devised a way to get rid of thermal noise, so we are stuck with it."

and random phase sine waves lumped together.

Now reconnect your antenna. Do you hear any additional hiss, pops or buzzing? In addition to thermal noise, you will have noise contributed by the amplifier circuit, and by RF noise from various natural and man-made sources in your environment, and of course any QRM present.

To convince yourself that there is natural RF competing with your signal, point your antenna at the sun—you will notice an increase of about 12 dB in noise!

You can measure the extra noise using an oscilloscope, or just your ear, to obtain the factor we need to add to thermal noise to get good answers for S/N (see sidebar).

Signal to Noise Ratio

We will continue our quest for a formula for S/N using just thermal noise for the time being, realizing that result will give us a S/N that is the theoretical limit.

Just as we could get the signal voltage from the received power, we can likewise solve for the noise voltage (again we will take the input impedance to be 50 ohms):

$$V_N = 8.9\times10^{-10}\sqrt{B} \qquad (4)$$

For example, if your receiver's SSB bandwidth is 3000 Hz, then V_N evaluates to about 1/20 microvolt.

We are now all set to combine the path loss equation (2) with the noise equation (4) to get a handy formula for voltage S/N:

$$S/N = \frac{V_R}{V_N} = 6.27\times10^{8}(l/r)\sqrt{P_T/B} \qquad (5)$$

So: given the transmitter power, the wavelength, the distance, and the signal bandwidth, we can compute the quality of our signal at the receiver!

Things we can learn from eqn (5)

- Although we used the impedance of 50 ohms in computing voltages, the resistance cancels out in the formula S/N. Therefore, *you cannot effect S/N by changing the antenna impedance or input impenance* of your receiver (as long as they are matched, of course).

- Choice of bands has a stronger effect than power or bandwidth. By moving from 20 meters to 40 meters, your S/N will double. You would have to quadruple your power to have the same effect!

- Using a narrow bandwidth communication mode, and corresponding narrow filters at the receiver will improve your signal as much as an increase in power. For example, switching from 3000Hz SSB voice communication to 30Hz PSK31, will boost your S/N ten times. You would have to increase your output power 100-fold to give the same result!

- The factor of P_T/B is power per Hz. Physicists call this “spectral brightness”. 100 watts in 3000 Hz, has the same brightness as 1 watt in 30Hz, and equal brightness gives equal S/N. No wonder that QRP people love CW and PSK31! When choosing a power and mode, remember that brightness (watts/Hz) is what counts.

Remember that eqn (5) gives us the voltage signal to noise ratio. That’s important because your receiver is a voltage amplifier, and what you see on the scope is certainly voltage. However, signal intelligibility (especially for digital modes), is closely related to the

Measuring Noise

With your receiver on, squelch open, and antenna disconnected, turn up your AF until you hear an audible hiss. If you have an oscilloscope, attach the probe to the external speaker, set the vertical gain to 100 mV/division, and turn up the AF until the jumble of waveforms spans approximately + and— one division.

Now attach the antenna. You will hear and see more noise. You might have to adjust the vertical gain, but eyeball the new voltage amplitude on the scope. Divide what you get by 100mV. Square the result to get the increase in noise power. Take 10 log() of that number, and you have your system’s noise in dB relative to thermal noise.

For the extremely noisy RF environment at my office, I found that at 14 MHz the audio noise voltage increased from 100 mV to 600mV, a factor of 6, so the noise power flowing into my receiver is 36 times thermal noise power, an increase of 14 dB.

power signal to noise ratio (your ears detect audio power more than amplitude). Since power is proportional to the square of voltage, we can just square the results of eqn(5), and we are done:

$$S / N = \frac{P_R}{P_N} = \left(\frac{V_R}{V_N} \right)^2 \qquad (6)$$

The rest of the story

So far we have computed the signal a receiver will detect based on line-of-sight geometry, isotropic antennas, and perfect thermal noise. It's now an easy matter to fine tune this result by including the real-world effects:

- An isotropic antenna radiates equally in all directions, but is physically impossible. Real antennas have gain or loss relative an isotropic radiator, which can be expressed in dBi. For example, dipoles have a gain of +2.1dBi, and so do verticals (over an ideal ground), while a rubber duck may introduce a relative loss. If we use a dipole on each end of the path, we pick up 4.2 dBi of gain which is over 60% improvement in power S/N. Of course a pair of Yagi's each with 10 dBi of gain and aimed at one another results in 100 times improvement in power S/N.

- Systems with feedlines will have some loss, expressable in dB. Sometimes there is no loss—two HT sets communicating in simplex have no feedline losses because there is no feedline. Usually however there will be feedline losses but they are quite small, especially for HF frequencies. Unless you are very fussy about getting precision answers, or there is a known and substantial feedline loss, there is little danger is considering the loss negligible.

- As discussed earlier, there may be additional noise to consider, which can be expressed in dB over thermal noise. Measuring the noise background at the receiver might be the only way to estimate this loss.

The assumption that will give us the most trouble is that the radiator is in free space and is in line of sight with the receiver. This is reasonably the case for FM simplex, repeater operations, all Part 15 devices, short range wireless devices, satellite communications, EME (moon bounce) and so on. HF communications over long paths involving sky waves, tropospheric ducting, and the like clearly involve a very different propagation mechanism.

In the next edition of *HamPost,* I am going to do a little analysis of how HF propagations works, both from theory and how to measure the loss using beacons, or known stations (such as WWV or ARRL code practice and bulletin sessions). For now we have to settle for a fudge factor of 45-55 dB of loss when we know propagation depends on skywave.

Putting this all together, we have:

$$\boxed{\begin{aligned} PL &= -22 + 20\log(\,l\,/\,r\,) + A_T - F_T \\ &\quad + A_R - F_R - HT \\ P_R &= P_T\,10^{PL/10} \\ P_N &= 4kTB\,10^{N_x/10} \\ S/N(power) &= P_R / P_N \\ S/N(voltage) &= \sqrt{S/N(power)} \end{aligned}}$$

where A_T and A_R are the antenna gains, F_T and F_R are the feedline losses, HT is the loss due to ionospheric correction, and N_X is the noise above thermal noise, all in dB.

The formulae above are easy enough to program into a calculator, or to set up in a spreadsheet, but to help you with doing your own calculations, you can download pathloss.exe, a free Windows program I wrote, available at www.softsci.com/hamradio/download/v1n1-pathloss.zip

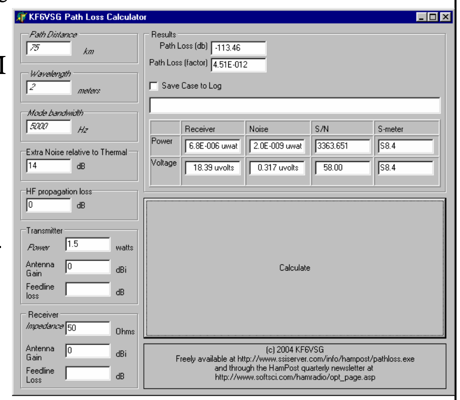


Figure 1. Pathloss.exe

No installation is necessary. Just launch the program, plug in the input numbers, and it does the rest.

Pathloss.exe is fun and interesting to play with. Here are some experiments you can program with it:

—WWV at 15MHz (20 m)

—Your HT to your local repeater

—Low earth orbit satellite (e.g., AO-27)

—Your cell phone

—An 802.11b wireless link

For next time

In the next issue of *HamPost*, we'll explore these cases and a few more, and then take what we learned and delve into the realm of S-meters and S-units, plus more on HF propagation.

End page

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