ARRL 100 YEARS

QEX

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November/December 2014

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A Forum for Communications Experimenters

Issue No. 287

KR4HO built this "Receiving Converter for 2 Meter Radios" to use with an old radio that covers the AM aircraft band as well as 2 Meter FM. Now he can tune that radio from 18 through 74 MHz and 218 through 235 MHz.

The national association for ARRL AMATEUR RADIO™
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Harold Kramer, WJ1B
Publisher

Larry Wolfgang, WR1B
Editor

Lori Weinberg, KB1EIB
Assistant Editor

Zack Lau, W1VT
Ray Mack, W5IFS
Contributing Editors

Production Department

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Publications Manager

Michelle Bloom, WB1ENT
Production Supervisor

Sue Fagan, KB1OKW
Graphic Design Supervisor

David Pingree, N1NAS
Senior Technical Illustrator

Brian Washing
Technical Illustrator

Advertising Information Contact:

Janet L. Rocco, W1JLR
Business Services
860-594-0203 – Direct
800-243-7768 – ARRL
860-594-4285 – Fax

Circulation Department

Cathy Stepina, *QEX Circulation*

Offices

225 Main St, Newington, CT 06111-1494 USA
Telephone: 860-594-0200
Fax: 860-594-0259 (24 hour direct line)
e-mail: **qex@arrl.org**

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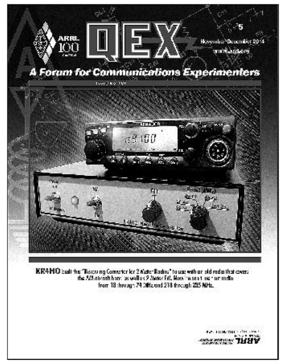
About the Cover

Colin Brackney, KR4HO, discovered an old Kenwood TM-241A 2 m FM transceiver that a friend had left with him. In addition to the 2 m band, this radio also tunes 118 to 135.995 MHz to cover the AM aircraft band. Colin decided to build a "A Receiving Converter for 2 Meter Radios" and can now receive AM and FM from 18 to 74 MHz. As a bonus, he added coverage for 218 to 235 MHz.



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The American Radio Relay League

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Officers

President: KAY C. CRAIGIE, N3KN
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Chief Executive Officer: DAVID SUMNER, K1ZZ

The purpose of *QEX* is to:

1) provide a medium for the exchange of ideas and information among Amateur Radio experimenters,

2) document advanced technical work in the Amateur Radio field, and

3) support efforts to advance the state of the Amateur Radio art.

All correspondence concerning *QEX* should be addressed to the American Radio Relay League, 225 Main Street, Newington, CT 06111 USA. Envelopes containing manuscripts and letters for publication in *QEX* should be marked Editor, *QEX*.

Both theoretical and practical technical articles are welcomed. Manuscripts should be submitted in word-processor format, if possible. We can redraw any figures as long as their content is clear.
Photos should be glossy, color or black-and-white prints of at least the size they are to appear in *QEX* or high-resolution digital images (300 dots per inch or higher at the printed size). Further information for authors can be found on the Web at **www.arrl.org/qex/** or by e-mail to **qex@arrl.org**.

Larry Wolfgang, WR1B

Empirical Outlook

The Year in Review

As we wrap up the November/December 2014 issue of *QEX*, I can't help but reflect a bit on the year that has been. Well, in reality I am quite happy that there are still nearly 3 months to go in 2014. I'm not ready for this year to be over just yet, but even from my early October perspective, it is rapidly coming to a close.

Of course everyone knows that 2014 has been the ARRL's 100th Anniversary. It has been a year-long celebration, with many memorable events. I hope many of our readers have enjoyed at least some of them.

As a ham for about 45 years, and also as a person who is very interested in the technical aspects of our hobby, sometimes my interest in on-the-air operating has waned a bit. I imagine that is fairly typical of anyone who has been licensed that long. I've never completely lost my interest in getting on the air, but sometimes it is easier to pick other pastimes. This year I have been more active throughout the year than I can remember in recent times. Several things have helped fuel that interest.

The ARRL Centennial QSO Party has been a lot of fun. I have enjoyed several multiple-hour operating stints of just getting on the air and chatting with fellow hams. There were a couple of "Red Badge Days" advertised, with an emphasis on getting on the air and contacting ARRL HQ Staff members as well as elected officials. Those activity days could only be successful with the ARRL Membership if the "Red Badge" folks were actually on the air making contacts. While I didn't spend any whole days operating, I did try to put in a couple of hours at a time, and I found this to be a lot of fun!

Another activity that has really captured my attention throughout the year has been the W1AW special event stations. I didn't put as much effort into contacting some of the earliest states, so I missed a few and am still waiting for the second go around for a couple of those. I won't miss them the second time. At this point I have contacted all but three of the states on both CW and SSB. I am feeling pretty confident about meeting my goal of a W1AW Worked All States certificate on both modes.

One unintended consequence of all this activity has been the realization that I was fairly close to earning several other awards. I have a mixed mode Worked All States Award going back to sometime in the early 1980s, and I also completed my mixed mode DXCC in 1995. I haven't actively pursued new band/mode states or countries towards endorsements since then. While looking at my Logbook of The World records late this Spring, I realized that I was not all that far from completing WAS and DXCC on both Phone and CW. Since then I've done a little targeted DX operating, and now have over 100 Phone QSL records in my LoTW account, and currently stand at 92 on CW. I've worked at least 8 more countries on CW, but the QSL records have not come through yet. With help from the ARRL Centennial QSO Party and the W1AW Special Event Stations, I have WAS confirmed on CW and Phone now.

As I looked at some other awards listings, I realized that I could also update my mixed mode Worked All Continents Award with both CW and Phone Awards. My plan is to apply for all of these awards/endorsements as soon as I have those last 8 DXCC CW QSL records in LoTW. This is really fun!

I have previously mentioned the ARRL National Convention in Hartford on this page. What a spectacular event that was! I have not been to a National Convention in quite a few years. This one would have been worth the trip from almost anywhere. I'm glad it was practically in my backyard!

Over the last 12 months I have been reading and learning more about software defined radio, and digging a little deeper into some of the math behind what goes on inside our modern radios. I have also found myself learning a little bit about *Linux*, or at least the *Raspbian* version used with Raspberry Pi computers. I'm still trying to gain a better understanding of why an operating system designed for use with this little computer always seems to require additions/deletions/other changes to make it work with various applications, but at least I have become a little familiar with the terminology. The Raspberry Pi is a fun little box to play with.

I have also recently mentioned that I have flashed the Broadband-Hamnet firmware into a pair of Linksys WRT54G wireless routers, to create my own little Mesh network. I still have a lot to learn about using this Mesh network, but it's a start. It looks like I'll have plenty of projects to keep myself busy for some time to come!

What gets you excited about Amateur Radio? I hope you have been enjoying and learning from the great articles that we have had in *QEX* over the past years. I can assure you that there are more good articles to come in 2015, but I also need you to write about the projects and topics that interest you. A lot of others want to learn from your experience and expertise! Have fun.

David L. Hershberger, W9GR

10373 Pine Flat Way, Nevada City, CA 95959; **w9gr@arrl.net**

Controlled Envelope Single Sideband

Introducing Controlled Envelope SSB; greatly increase your SSB "talk power" by accurately limiting envelope peaks in the SSB modulator. Generate SSB without the big overshoot peaks that make ALC necessary with conventional SSB modulators. Watch your wattmeter read higher than before.

Abstract

Achieving simultaneous accurate control of both amplitude and bandwidth is a difficult problem. When amplitude-limited audio is filtered to limit its bandwidth, the filter may overshoot substantially. It loses its amplitude limiting ability. If the resulting overshoots are clipped, the amplitude is controlled but the signal's bandwidth increases because of the clipping distortion. The signal loses its bandwidth limiting. Systems exist for correcting audio low-pass filter overshoot. But single sideband (SSB) is a more difficult problem, because of the inevitable Hilbert transform regardless of the method of SSB generation. ALC systems reduce the amplitude of an SSB signal in response to overshooting envelope peaks. Fast ALC may result in clipping and splatter. Slow ALC will significantly reduce transmitted power. This paper presents a method for generating SSB without system overshoots. The result is higher transmitted power without audible distortion.

Objective

One of the reasons for the existence of the Amateur Radio Service is the development of new techniques for radio communication. I hope that this paper will present a useful method for improving the effectiveness of SSB transmitters for both amateur and commercial applications. This technique is being placed into the public domain and in particular, the "ham domain."

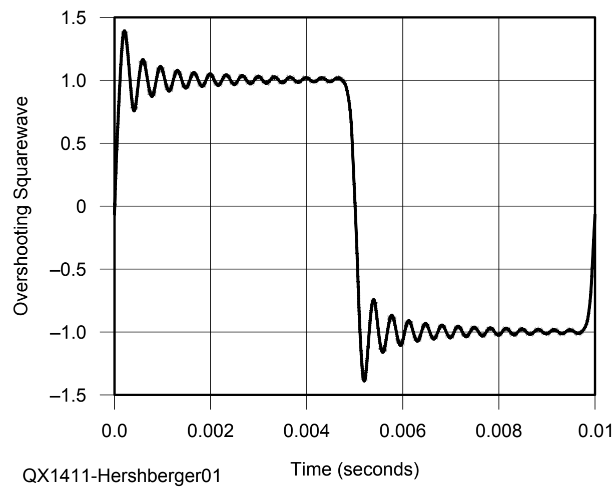


Figure 1 — 100 Hz square wave filtered by 3 kHz elliptic low-pass filter.

Benjamin Franklin expressed it well: "As we enjoy great advantages from the inventions of others, we should be glad of an opportunity to serve others by any invention of ours; and this we should do freely and generously."

Background

In the 1970s, FM stereo broadcasters' on-air loudness was affected by overshoot in the sharp cutoff 15 kHz low-pass filters (or 19 kHz notch filters) used in stereo generators of the time. These filters were necessary to prevent crosstalk between the left plus right and left minus right subchannels. As the stereo generator's filters would overshoot, the modulation would have to be reduced to keep infrequent peaks from exceeding 100% modulation. A number of manufacturers responded with systems that controlled overshoot accurately, allowing a loudness increase of 2 to 4 dB without any audible increase in distortion. The overshoots themselves contained relatively low energy. Eliminating them resulted in a substantial loudness increase without any perceptible increase in distortion. Such overshoot control systems are still in widespread use today.

In this paper a classic method for real signal low-pass filter overshoot control is presented. Then the technique is extended to the complex baseband signals used to generate SSB.

Overshoot Control for Low-pass Filters

The problem with low-pass filters is that they overshoot on amplitude-limited signals such as square waves. Figure 1 shows what happens when a 100 Hz square wave signal, accurately limited to a value of 1.0, is filtered by a sharp cutoff 3 kHz elliptic low-pass filter. It overshoots substantially.

There are several classes of non-overshooting or low-overshoot low-pass filters. These include Gaussian, Bessel, transitional, and parabolic filters. These are all linear filters. All of them have a roll-off characteristic that is far too slow to be useful for FM broadcasting or SSB generation, however. Therefore, an effective system for overshoot control must be nonlinear.

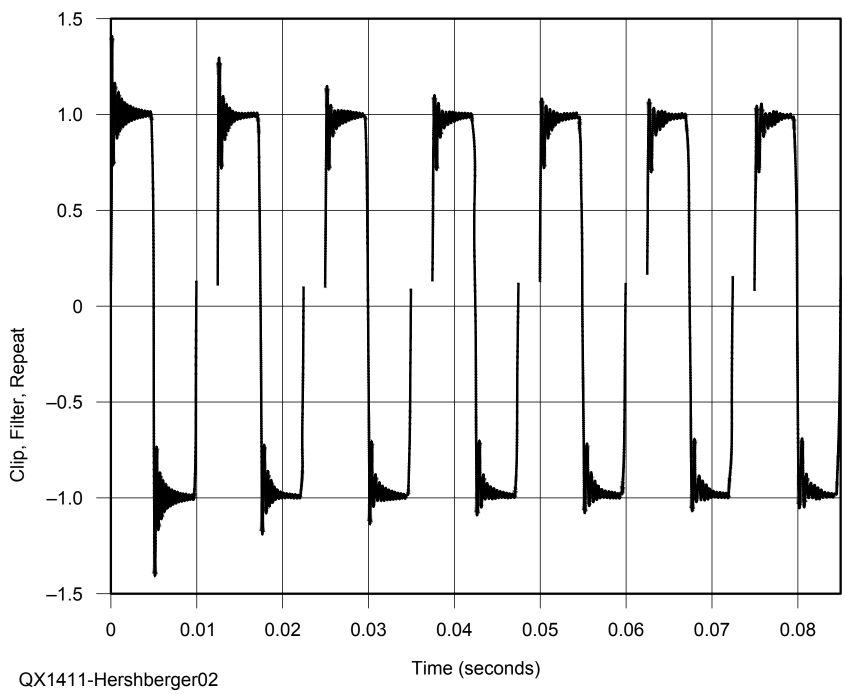


Figure 2 — Effects of repeated clipping and filtering.

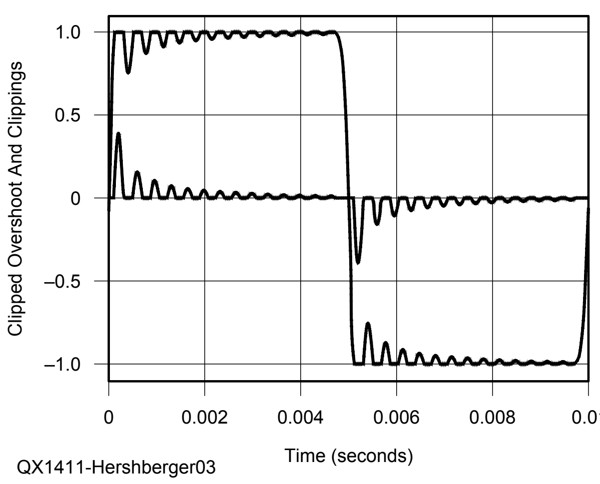


Figure 3 — Overshoots clipped off and separated.

My system for overshoot control of low-pass filters (expired US patent #4,134,074) started from the notion that overshoot could be reduced, but not eliminated, by clipping and refiltering. Clipping controlled the overshoots, but created harmonic distortion beyond the cutoff frequency of the filter. A filter after the clipper would remove the harmonic distortion — but it would overshoot. The overshoot of the second filter was not as high as the first filter, however. Theoretically, the clip, filter, clip, filter process could be repeated *ad nauseum* until the overshoot was insignificant. Figure 2 shows what happens with seven stages of filter, clip, filter, clip, and so on.

One cycle of a square wave is shown after the first, second, third stage, and so on. Even after the seventh pass, the overshoot is still over 4%. This method does not converge very quickly. Such a system would be impractical in analog circuitry. So, I came up with a method to make the system converge in “one fell swoop.” What was needed was something that did “more than clipping.”

A clipper can be visualized as a circuit that creates a series of peaks, which are then subtracted from the input signal. That is, it creates “clippings” that are the “tops” (and bottoms) of the input waveform. These are subtracted from the input to create a clipped waveform. The gain of the “clippings” in a simple clipper is unity.

Figure 3 shows what happens when the overshoots are simply clipped off. The overshoot “clippings” (lower amplitude trace) may be separated by subtracting the output from the clipper from its input. When the square wave with clipped overshoots is filtered, the second waveform in Figure 2 above is the result. The overshoot is reduced but not eliminated.

Something that would do “more than clipping” could apply a gain factor to the “clippings” — such as a gain of 2. The original overshooting signal shown in Figure 1 with such processing is shown in Figure 4.

Note that the overshoots that would have exceeded 1.0 are turned around and made to go the other way. The signal of Figure 4 contains out of band distortion and should be low-pass filtered with linear phase. When the “more than a clipper” processes the signal, and the result is linear phase low-pass filtered, the result is a large reduction in overshoot. Almost all of the overshoot is removed in this single more-than-clip and filtering process. The result is shown in Figure 5. The block diagram for this system is shown in Figure 6.

Audio is assumed to come from a peak limiter device, which ensures that nothing over 100% is being applied to the system.

This signal is low-pass filtered, and the low-pass filter will overshoot. The output of the low-pass filter is clipped. The output of the clipper is subtracted from the clipper's input, to obtain the "clippings." The clippings are then amplified by a factor of about 2. They are then subtracted from the low-pass filter's output to create the waveform of Figure 4. That waveform is then low-pass filtered again with a linear phase low-pass filter. The result is simultaneous control of both peak amplitude and bandwidth.

There are two possible enhancements to this system. First, the simple gain factor of 2 generally results in insufficient overshoot correction at lower amplitudes, and too much correction at full amplitude. So rather than applying a simple gain factor of 2 in the "gain" element shown in Figure 6, the clippings are divided by the greater of 1.0 or the absolute value of the clipper input. Then a slightly lower gain factor of 1.9 is applied. The signal shown in Figure 4 has been processed this way. As an expression, the correction is:

$$corr(t) = \frac{1.9 \bullet clippings(t)}{\max\left[1, \left|lpf(t)\right|\right]} \qquad \text{[Eq 1]}$$

Where:
corr is the additive correction signal
clippings is the input of the clipper minus its output
max is the function that returns the greater of its two inputs
lpf is the output of the first low-pass filter
The vertical bars operator is absolute value

Second, there is a way to reduce the signal processing complexity. Rather than passing the entire signal through the second low-pass filter, just the "clippings" may be passed through the filter if the rest of the signal is delayed to match the delay of the second low-pass filter. This modification is shown later in Figure 13.

When we generate SSB, the system as described above will not work for overshoot control. In DSP, single sideband is usually generated by creating two orthogonal baseband audio signals, such as the well-known phasing method. The phasing method as implemented in analog hardware required very close matching and phasing of the two signal paths. This was difficult to do in analog

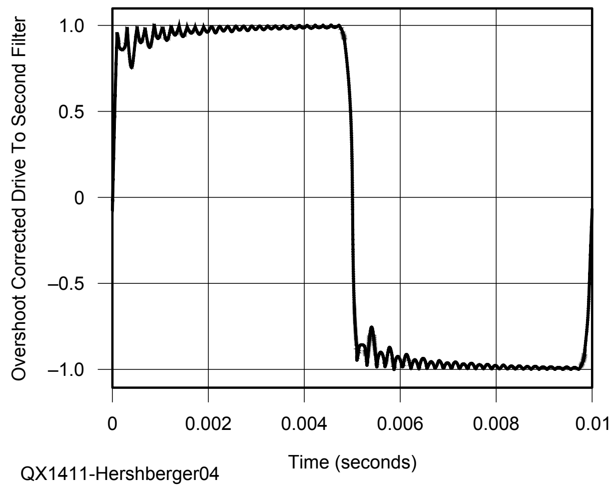


Figure 4 — 100 Hz filtered square wave processed with "more than clipping."

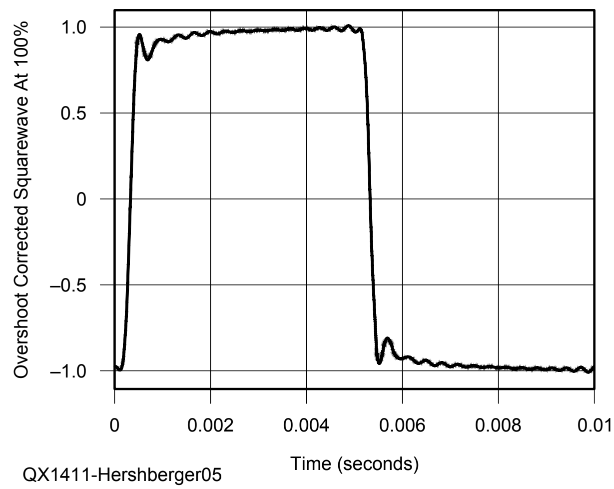


Figure 5 — Square wave filtered with overshoot compensation.

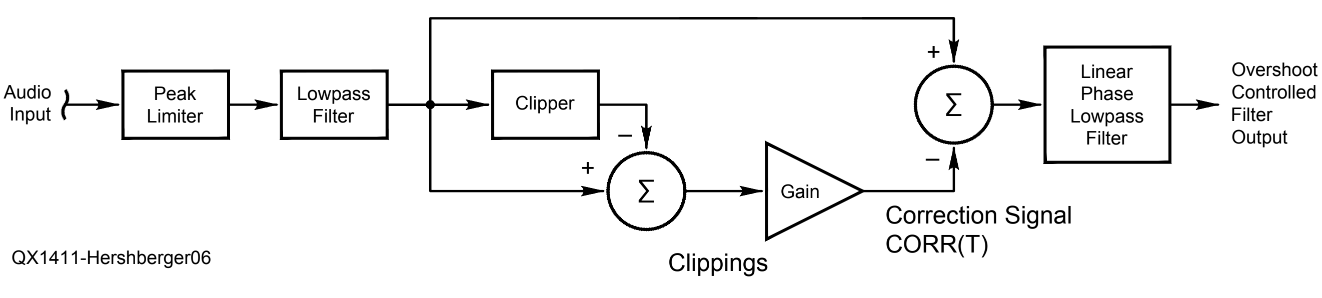


Figure 6 — Overshoot compensation system for real signals.

Figure 7 — First three Fourier components of a 100 Hz square wave. Notice that zero crossings are coincident at 0, 0.005, and 0.01 seconds.

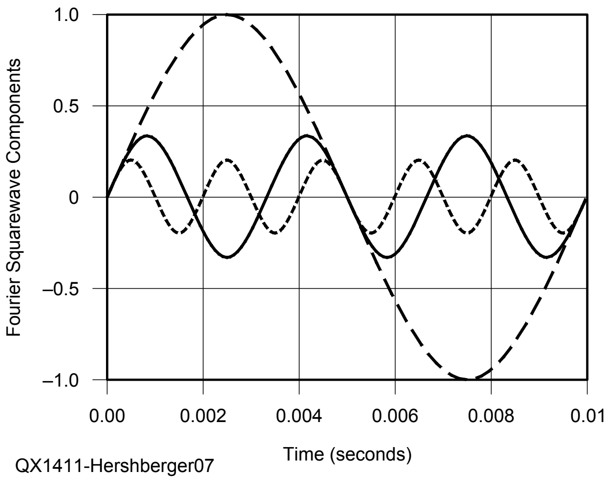
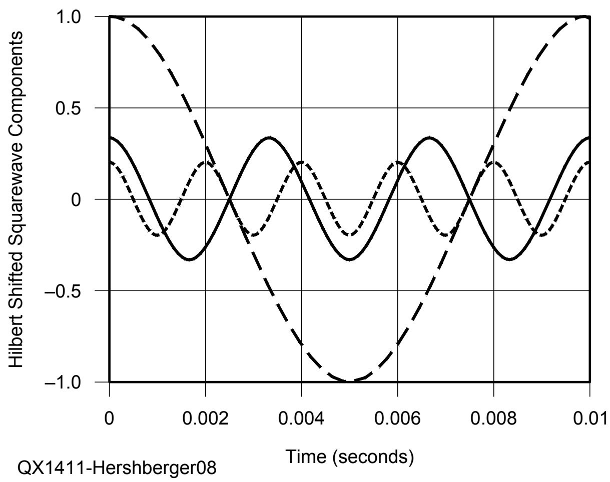


Figure 8 — First three Fourier components of a 100 Hz square wave, all shifted 90° (Hilbert transformed). Notice that waveform peaks, rather than zero crossings, are coincident at 0, 0.005, and 0.1 seconds.



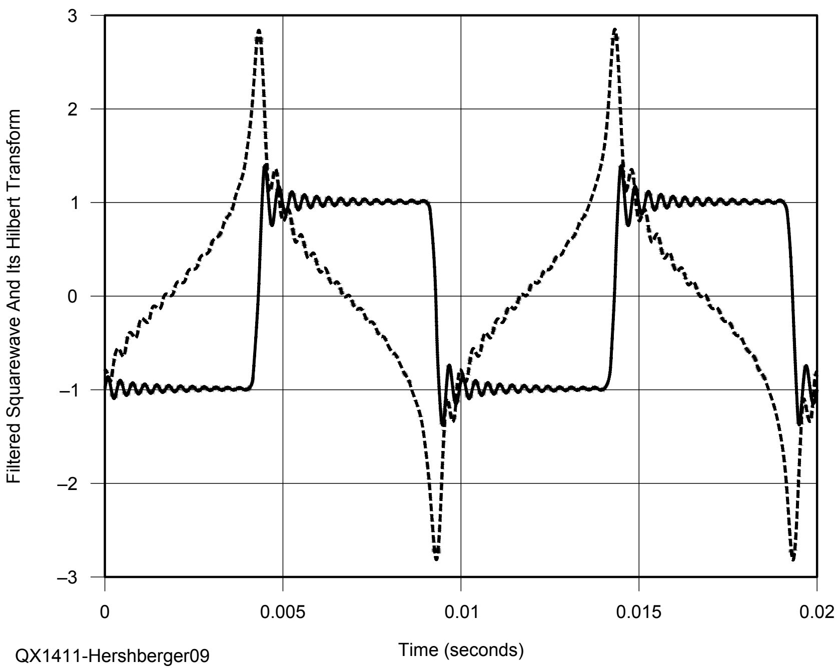


Figure 9 — Filtered square wave and its Hilbert transform.

hardware but exact matching is trivial in DSP. If we try to control overshoot independently in the two paths, however, the system will not work. A different approach is required for complex signals.

Overshoot Control for Complex Signals

Now let us extend the overshoot correction system for real signals to complex signals, so that we can generate SSB without overshoots.

What is a complex signal? It is a signal with an in-phase (I) part and a quadrature (Q) part, which can also be treated as real and imaginary parts. Physically, there are two signal paths. Complex signals are also known as analytic signals.

The phasing method for generating single sideband is an example of creating a complex signal in physical hardware. We take a modulating signal (speech audio) and shift all of its frequency components by 90° (a Hilbert transform). The original audio is applied to an RF mixer operating at one phase, and the phase shifted audio is applied to another RF mixer operating with a 90° RF phase shift. When the two double sideband signals are added, one sideband cancels and the other adds. We have single sideband.

$$ssb(t) = m(t)\cos(\omega_c t) + H\left[m(t)\right]\sin(\omega_c t)$$

[Eq 2]

Where:
ssb(t) is our single sideband signal
m(t) is the modulating function
H( ) indicates the Hilbert transform
ω_c is the radian carrier frequency, and
t is time.

In modern DSP based modulators for many signal types, including 8 level vestigial sideband digital television, orthogonal frequency division multiplexing (OFDM) and other types, the signals are generated at baseband and then up converted to the operating frequency in the later stages. Overshoot compensation of SSB signals is also best done at baseband.

Regardless of how we generate SSB (filter method, phasing method, Hilbert transform method, or Weaver method), we always get a signal that has a Hilbert transform in it.[1] Choose any demodulation phase for an SSB signal, and if you shift that RF phase 90°, the demodulated audio phase will also shift 90° for all frequencies. That audio phase shift constitutes a Hilbert transform.

A square wave causes large peaks when Hilbert transformed. A square wave is a summation of a fundamental sine wave plus odd harmonic sine waves, each diminished by the harmonic order (third harmonic at

[1]Notes appear on page 13.

1/3, fifth harmonic at 1/5, and so on). All of the sine waves have their zero crossings coincident. This is shown in Figure 7.

Now if we shift all of the audio frequencies by 90°, those sine waves turn into cosines. While the sine wave components of the square wave share zero crossings, the cosine components are all coincident at their peak amplitudes! This is shown in Figure 8.

When we apply a Hilbert transform to baseband audio, we can get some really nasty waveforms. Beginning with our low-pass filtered 100 Hz square wave example, Figure 9 shows what the Hilbert transform looks like. The Hilbert transform overshoots are much more severe than the peaks of the filtered square wave. The envelope of an SSB signal generated from this Hilbert transform pair will spend most of its time overshooting. This is why the Hilbert transform of a square wave is so spiky looking. And for that reason, clipped speech applied to an SSB modulator will also produce some large spikes in the RF envelope.

In Figure 10, 1.0 is "full modulation," corresponding to an input signal amplitude of 1.0. The envelope is much larger, creating some nasty peaks. In fact, the envelope rarely dips below 1.0 in Figure 10!

Again, this envelope (or something even worse) will be produced regardless of the manner of SSB generation (filter, phasing, Hilbert, or Weaver), because I and Q are always related by a Hilbert transform in an SSB signal.

For a complex signal, what we are ultimately trying to control is not the amplitude of I and Q individually. We are trying to control the envelope of the resulting signal, which is the modulus, or

$$\sqrt{I^2(t)+Q^2(t)}.$$

To convert the real overshoot control system to complex, we must form the envelope signal and use it to control the gain of both I and Q. For this conversion, we should note that a clipping function can also be done by a divider. If we divide a real signal by the maximum of 1 or the absolute value of the real signal, the real signal will be clipped. It is this approach (division) that will extend the real technique to complex signals. Instead of dividing by the absolute value of the real signal, we will divide by the modulus of the complex signal.

The complex signal overshoot control system with a Weaver SSB modulator is shown in Figure 11. Most of the signal paths shown in this control system are complex. The wide lines represent complex signals and the narrow lines are real signals.

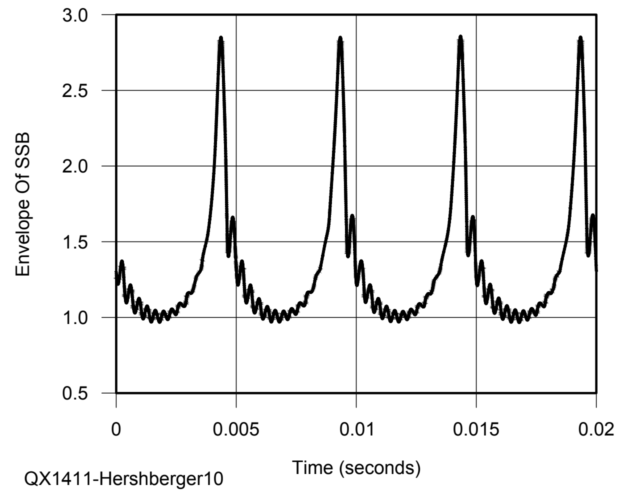


Figure 10 — Envelope of square wave SSB from I and Q signals of Figure 9.

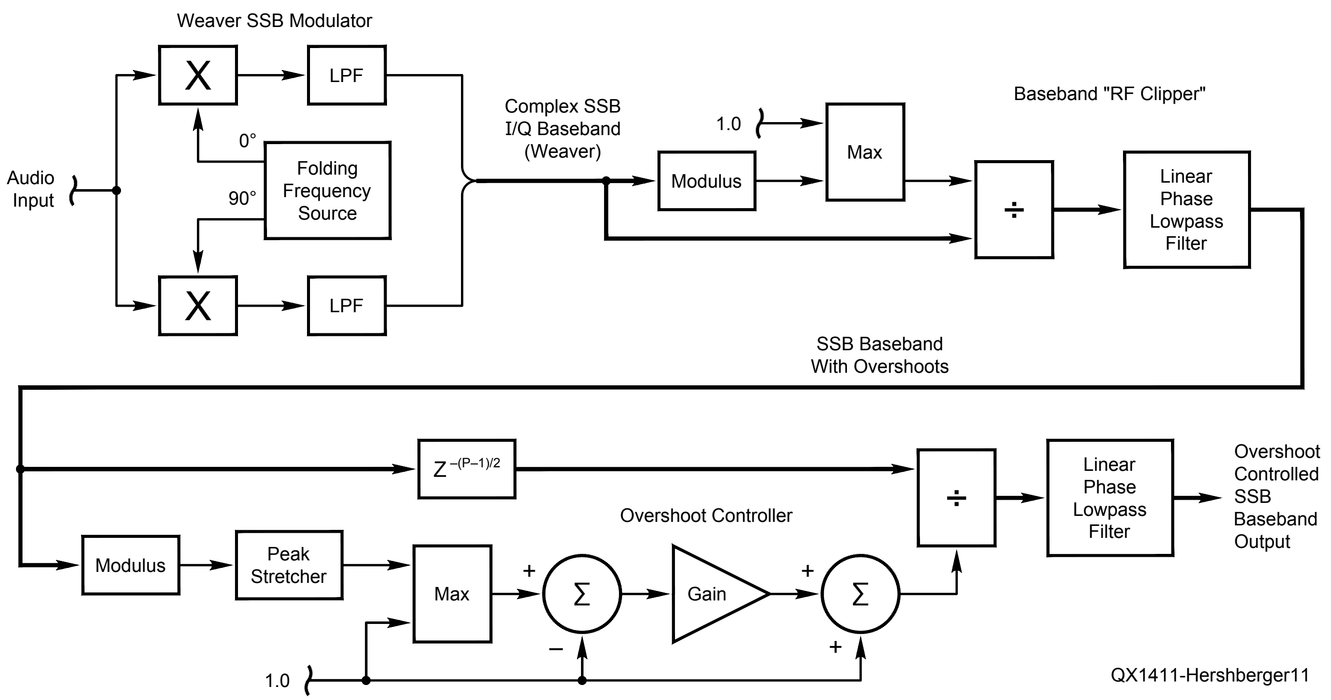


Figure 11 — Overshoot control for single sideband.

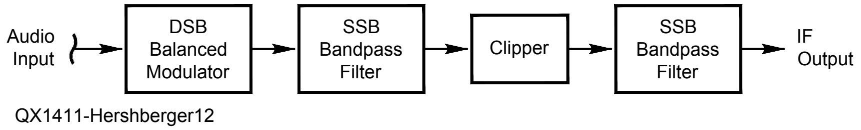


Figure 12 — Conventional analog RF clipper. RF cycle overshoot, spectral truncation, and filter group delay effects reduce its effectiveness.

Overshoot control for single sideband is simplest if we use the Weaver system (see the Generating SSB with Complex Math sidebar) because the filters operating on the real I and Q signals are all real (no complex coefficients). A further advantage of the Weaver system is that the bandwidth of the baseband I and Q signals is less than half what it would be with the Hilbert transform method. This allows use of a lower sampling rate. If the sampling rate is cut in half, the number of filter coefficients will also be cut in half for the same shape factor. This results in a Weaver SSB computation rate ¼ of what it would be with a Hilbert transform or phasing method SSB modulator.

Figure 11 shows a Weaver modulator for generation of SSB baseband signals from incoming peak-limited audio. The folding frequency for a 300 to 3000 Hz system would operate at 1650 Hz. The two low-pass filters have a bandwidth of 1350 Hz. The resulting baseband spectra are folded through DC.

The subsequent overshoot control system is divided into two subsystems. First there is a baseband "RF clipper." It operates at audio baseband, and not at RF, but its operation is similar in function to a conventional "RF clipper." See Figure 12. The purpose of this function is to remove most of the overshoot from the process of generating SSB. *This is not intended to be an audio processor*. Clipping threshold is set to 0 dB, which is referred to an audio tone at maximum input level.

"RF clipping" is interesting because it results in no harmonic distortion — only intermodulation distortion. As such it is usually preferable to simple audio baseband clipping, which produces both harmonic and intermodulation distortion. "RF clipping" of a single sine wave tone results in no distortion at all.

A conventional analog "RF clipper" generates single sideband at some nonzero frequency, clips the RF cycles, then band-pass filters the result to eliminate out of band distortion components. The result is an SSB signal with greater average power. A conventional RF clipper cannot control peaks very well, however. It controls instantaneous peak RF amplitude, which is not the same thing as peak RF *envelope* amplitude. As each RF cycle is clipped, the clipped sine waves may be approximated as trapezoidal waves. RF harmonics are produced. In the limit (a large amount of clipping) the waveform will approach a square wave. When a square wave is filtered back to sine wave, the result will be overshoots. The peak amplitude of the fundamental component of a square wave is larger than the square wave peak by $4/\pi$. That would be 27% overshoot. Additional overshoot results from spectral truncation and group delay distortion in the second filter.

The baseband "RF clipper" shown in Figure 11 does not limit the instantaneous amplitude of RF cycles, because there are no RF cycles. Instead it limits the RF envelope. The RF envelope is the modulus of the complex time domain signal. There will be no $4/\pi$ overshoots. There will, however, still be some overshoot from spectrum truncation.

The modulus (envelope) signal is first generated. The larger of either the envelope or 1.0 is applied to the denominator input of a complex divider. The I and Q signals are divided by this inverse gain value. The modulus of the divider output will be perfectly limited to 1.0. The nonlinear processing will result in some out of band distortion products, however. The distortion is removed by a complex low-pass filter (which is really just two real low-pass filters for the Weaver system). The output of the low-pass filter will have some overshoots.

The overshoots from the RF clipper are then processed in the next subsystem shown in Figure 11. Again, the modulus is formed and the maximum of the modulus or 1.0 is generated.

The modulus value is next applied to a "peak stretcher." This simply takes the maximum of the current sample and several previous samples. The number of samples will depend on the sampling rate. The recommended length of the peak stretch window is given by Equation 3.

$$Pwin \approx \frac{0.3}{BW} \quad \text{[Eq 3]}$$

Where:

Pwin is the length of the peak stretch window in seconds (*Pwin* is approximately 111 μs), and *BW* is the bandwidth of the SSB signal (2700 Hz in this case).

Equation 4 converts this to samples.

Generating SSB with Complex Math

Let's look at the different ways to generate single sideband from a complex math perspective. There are three commonly used ways to generate SSB: the filter method, the Hilbert transform method (a variant of which is known as the "phasing method") and the Weaver method (also known as the "third method"). In analog circuitry the filter method is most common. In DSP, the Hilbert and Weaver methods are more common. In DSP, obtaining the two perfectly matched signal paths is trivial for Hilbert and Weaver, but generating a high Q band-pass filter for the filter method at high sampling rates is difficult.

First we look at the filter method. The filter method uses no complex math at all. All of the calculations are real. Nevertheless, we will show the mirrored positive and negative spectral components as part of Figure A.

The top line of Figure A shows the spectrum of the baseband modulating signal. Next is the carrier signal. The third spectrum shows the output of a balanced modulator, which multiplies the audio and the carrier. We see a double sideband signal at positive and negative frequencies. The next step is to apply a band-pass filter that passes the upper sideband and rejects the lower sideband. The final spectrum is our real SSB signal.

Next we will generate SSB with the Hilbert transform method. This will get us into some complex math. Figure B shows the progression of the generation of the signal.

The top spectrum is the modulating audio. The next spectrum is the response of a filter derived from a Hilbert transform. *H* is a Hilbert transform multiplied by $-j$, plus a unit real impulse response with a time delay of one half of the Hilbert transform delay. This sum of a Hilbert transform in the imaginary coefficients plus the delayed real unit impulse will suppress the negative frequencies, but allow the positive frequencies to remain.

The next spectrum shows the result of filtering *m(t)* with *G*. Only positive baseband frequencies remain. The next spectrum is a complex carrier sinusoid, at a positive frequency only. The next spectrum shows what happens when the complex positive frequency audio is multiplied by the positive carrier. Now we have complex SSB, at a positive frequency only. Finally, we discard the imaginary part of the signal. The result is a real SSB signal with mirrored positive and negative frequency components.

A variation on the Hilbert transform

method uses phase difference networks. Analog phasing method SSB exciters used phase difference networks rather than Hilbert transforms. A set of phase difference networks maintains a 90° audio phase shift between its outputs, but it incurs additional phase distortion beyond that of a Hilbert transform system. Phase difference networks are easier to implement in analog circuitry. In DSP, fewer calculations are required for phase difference networks if an infinite impulse response (IIR) implementation is used.[5]

Finally we will generate SSB using the Weaver method. Figure C shows the spectrum at each step in the process of generating the signal.

As before, the top spectrum is the modulating baseband audio. The next spectrum is a complex frequency-shifting or "folding" carrier, selected to be in the middle of the audio passband. The third spectrum shows what happens when the baseband audio is multiplied by the shifting carrier. The mirrored spectrum is shifted downwards. The next spectrum shows the response of a real low-pass filter. It is symmetrical around DC. In actual implementation, it consists of two identical real filters, one filtering the real part and the other filtering the imaginary part. After this filter is applied to the shifted baseband audio, what remains is the single audio spectrum folded through DC. Next we can multiply that complex folded audio spectrum by a complex carrier. The result is a complex (positive frequency only) single sideband signal. Finally, we discard the imaginary part, leaving a real single sideband signal with mirrored positive and negative spectral components.

In this article, we assume that the Weaver method is being used to produce single sideband. Since the overshoot correction process requires nonlinear processing followed by additional filtering, that filtering will be simplest to implement with real filters (no complex coefficients) of the type used in the Weaver method. Overshoot control can be done with the Hilbert method, but the filters must all be complex instead of real. Alternatively, since the Hilbert or phasing method baseband signals are already in analytic form, they may be frequency shifted downward with a complex multiplication by a complex tone to put them in spectrally folded Weaver format. After the overshoot correction is completed (using real filters) the baseband signals can be shifted back to Hilbert format by multiplying by the conjugate of the same complex tone. Note that with analytic signals, no filtering is required because there are no unwanted spectral images.

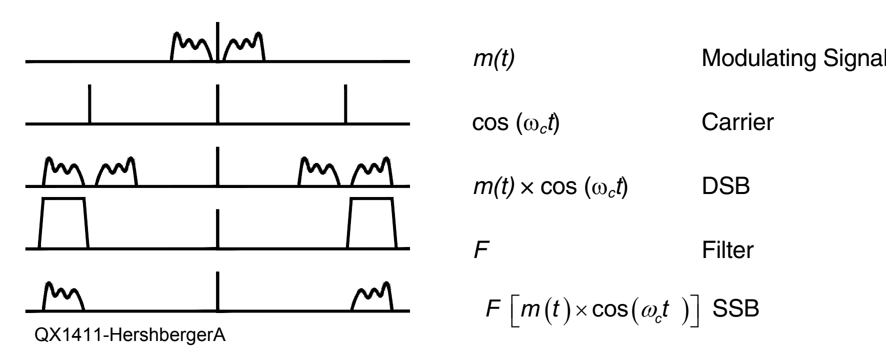


Figure A

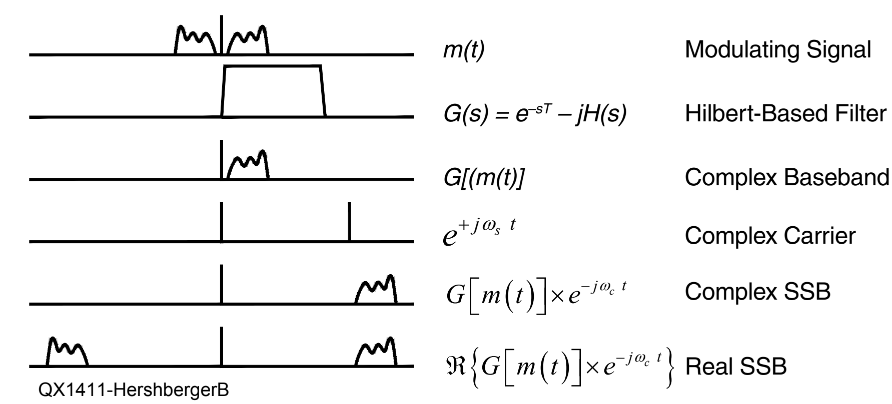


Figure B

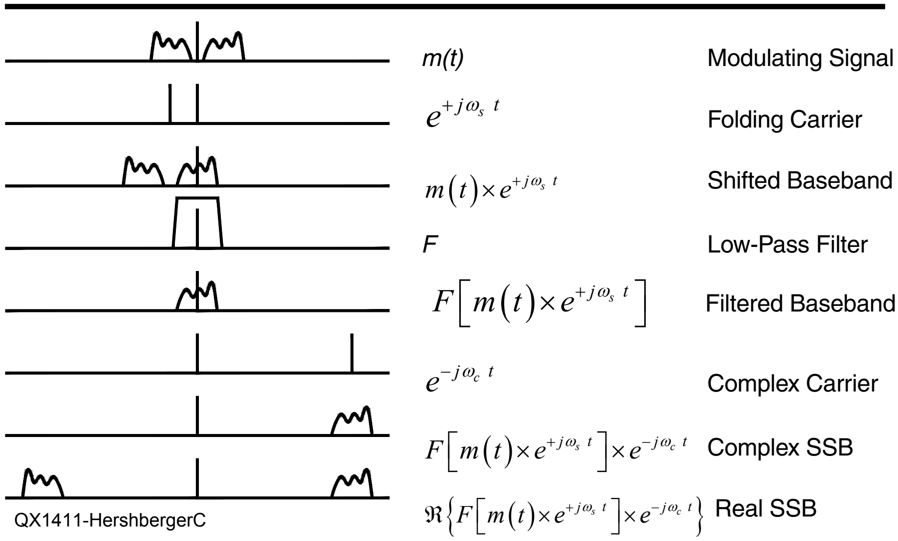


Figure C

$$Pn = max\left[3,\ round\left(Pwin \cdot Fs\right)\right]$$

[Eq 4]

Where:
Pn is the number of samples
max( ) selects the largest value from its arguments
round( ) rounds to the closest integer, and
Fs is the sampling rate in Hz.
Pn should be an odd number and at least 3. For example, for a 2.7 kHz bandwidth and a 48 kHz sampling rate, the closest integer value for *Pn* would be 5. Note that there is a compensating delay in the signal path such that the peak is stretched equally to both preceding and subsequent samples. The compensating delay for $p = 5$ would be 2 samples. Odd values of *Pn* are suggested so that the compensating delay will have an integer number of samples.

The peak stretcher may be omitted. Without it the overshoot will be about 5% on voice peaks. With it, the overshoot is reduced to about 1.5%.

The output from the peak stretcher has unity subtracted from it to produce output only when there is overshoot. The resulting signal is applied to a gain factor of approximately 2.0. With low sampling rates and a peak stretch of 3 samples, the optimum gain factor could be less than 2.0. After 1 is added to return the quiescent gain to unity, it is applied to the denominator input of a complex divider. The divider will perform a "more than clipping" function similar to what was done in the real overshoot compensation system. The difference is that the correction is applied in a divider rather than additively, and the gain control affects the real and imaginary components equally.

Finally, the overshoot compensated complex baseband signal is low-pass filtered to produce CESSB. If the final filter is made slightly wider in bandwidth than the incoming signal, there will be less overshoot. The Weaver filters used for SSB generation had a cutoff frequency of 1350 Hz. The final filter has a bandwidth of 1450 Hz. The same 1350 Hz filter could have been used with slightly increased overshoot (a small fraction of 1%).

Note that the entire system of Figure 11 has no effect on signals below 100% modulation. Applying a steady state tone below 100% will not invoke operation of any of the clippers. The system only affects transients that would produce overshoot. Audio compression to increase density, compression, peak limiting, and so on should be applied before the audio input to Figure 11.

DSP Economy

It is possible to reduce the amount of DSP horsepower required to control overshoot by implementing the filtering method shown in Figure 13. The method is shown for the baseband "RF clipper" but it may also be applied to the overshoot correction processing block as well.

In Figure 13, the input to the divider is subtracted from its output. The result is the nonlinear correction signal that has been added to reduce the envelope peaks. Rather than pass the entire signal through the low-pass filter, only the correction signal is passed through the filter. Since the correction signal has low energy, the specifications of the linear phase low-pass filter may be relaxed in both the passband and the stopband. With fewer filter coefficients than would otherwise be required, the overall calculation rate is reduced. Notice that the main signal path only incurs delay and no filtering. If the low-pass filter is a finite impulse response (FIR) filter, it should have an odd number of coefficients so that the delay line may have an integer number of sample delays.

Sampling rates should be mentioned at this point. Generating the initial SSB signal requires only that the sampling rate should be higher than twice the bandwidth of the baseband signals. The Weaver SSB generation method has a big advantage here since a typical 2.7 kHz wide SSB signal only requires a 1350 Hz baseband bandwidth. (The Weaver baseband spectrum is folded through DC.) The sampling rate could be as low as 3 to 4 kHz. The "RF clipper" and the overshoot compensator both generate wider bandwidth correction signals, however. To avoid aliasing, the sampling rate at that point should be higher. The simulations for this article used a sampling rate of 48 kHz for all processing, which is overkill. Inspecting the spectra of the nonlinearly processed Weaver baseband signals show that the bandwidth grows to only about 5 kHz at the –60 dB point. So the sampling rate for the nonlinear processing could be as low as 10 to 12 kHz, provided that the Weaver method is used.

In practice, input audio filtering and Weaver SSB generation could be done at a sampling rate of 8 kHz. Then a half band interpolator could raise the sampling rate to 16 kHz where all subsequent nonlinear processing could be done.

Results

A test audio file was prepared using SOX, an open-source command line audio processing program.[2] SOX is an acronym for "SOund eXchange." The SOX code used applies filtering, fast gain control, and peak limiting to a test audio file of my speech. The processing is aggressive, producing 100% peaks frequently.

Simulation was done using GNU *Octave*, an open-source matrix math program.[3] *Octave* reads in the processed audio file, and appends to it a 1 kHz tone at 100% modulation as an amplitude reference. (Here "100%" corresponds to the maximum PEP capability of the downstream RF power amplifier.) SSB is generated using the Weaver method. The envelope of the SSB generator is shown in Figure 14.

The envelope is normalized to 1.0, which corresponds to the peak of the audio signal.

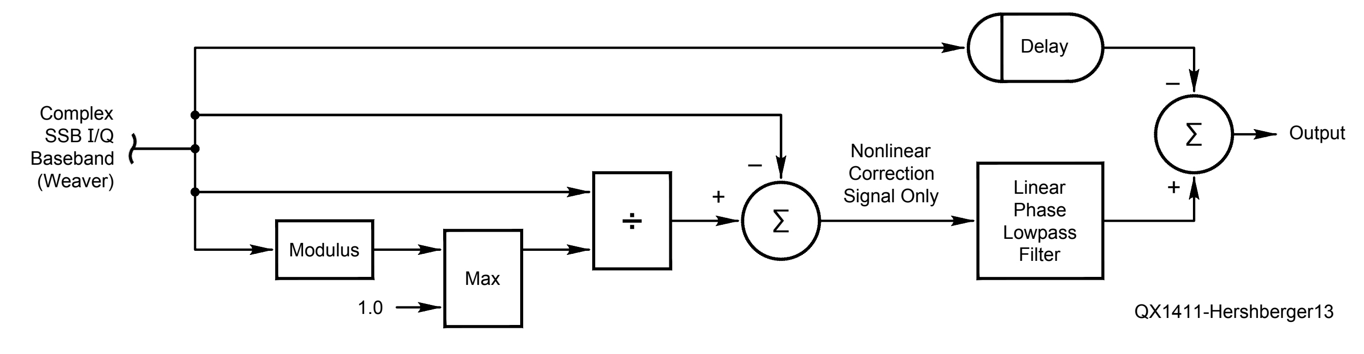


Figure 13 — Method for reducing calculation rate.

Overshoot is 59%. The tone burst at the end has overshoots at its beginning and end, but the steady state portion is at the nominal 1.0 level.

Figure 15 shows the envelope after the baseband "RF clipper" stage. Because of spectral truncation (filtering), the RF envelope is still not controlled. It is better than the conventional SSB generator, but there is still 21% peak overshoot with this audio sample. (A conventional analog RF clipper such as the one shown in Figure 12 would have even worse overshoot, because of the $4/\pi$ problem and nonlinear phase filters.)

Figure 16 shows the final output from the system. The peaks are well controlled. Overshoot is only 1.6%. The overshoots at the beginning and end of the tone burst are gone too.

Several supporting files have been placed on the ARRL *QEX* files website.[4] Included are:

1) The original unprocessed audio WAV file (SSB-test-wideband.wav).

2) The SOX command script that filters the input audio to SSB bandwidth, then compresses, and limits the audio (Sox-ssb-process.bat).

3) The FIR filter that restricts the audio bandwidth to 300 to 3000 Hz (SSBBPF.TXT). This filter is used by SOX.

4) The WAV file of the SOX-processed audio (SSBaudioprocessed.wav). This file was used to create Figures 14, 15, and 16.

5) GNU *Octave* source code which generates SSB, "RF clipped" SSB, and overshoot compensated SSB (cessb.m). A version compatible with *Matlab*™ is also included (cessb_matlab.m).

6. The demodulated audio output WAV file from the Weaver SSB generator signal of Figure 14 (ssbdemod.wav).

7. The demodulated audio output WAV file from Figure 16 (cessbdemod.wav).

The last two audio files can be compared to hear how using SSB peak control has very little effect on audio quality.

Implementation

Implementation can be done efficiently with a DSP chip and/or FPGA code. This method of SSB generation is something best integrated into a transmitter or transceiver, as opposed to an external add-on box. An add-on box would have to generate SSB at the radio's intermediate frequency and inject it in place of the radio's original SSB modulator.

System setup is best done with a single reference sine wave tone at 100% modulation. The clippers should be set to operate on anything that exceeds the envelope of the reference tone. If the incoming signal is low

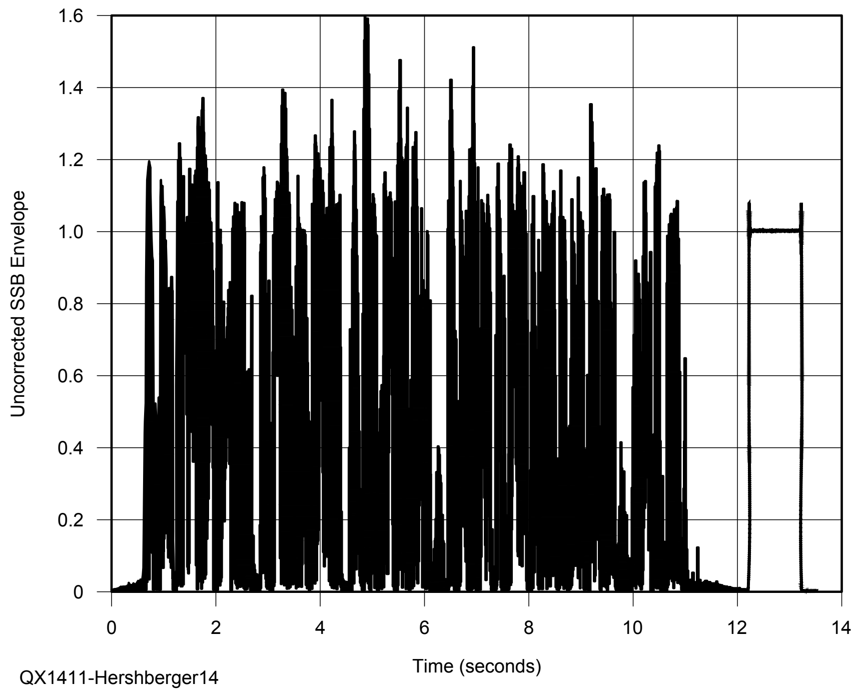


Figure 14 — Envelope of conventional SSB generator. Overshoot is 59%.

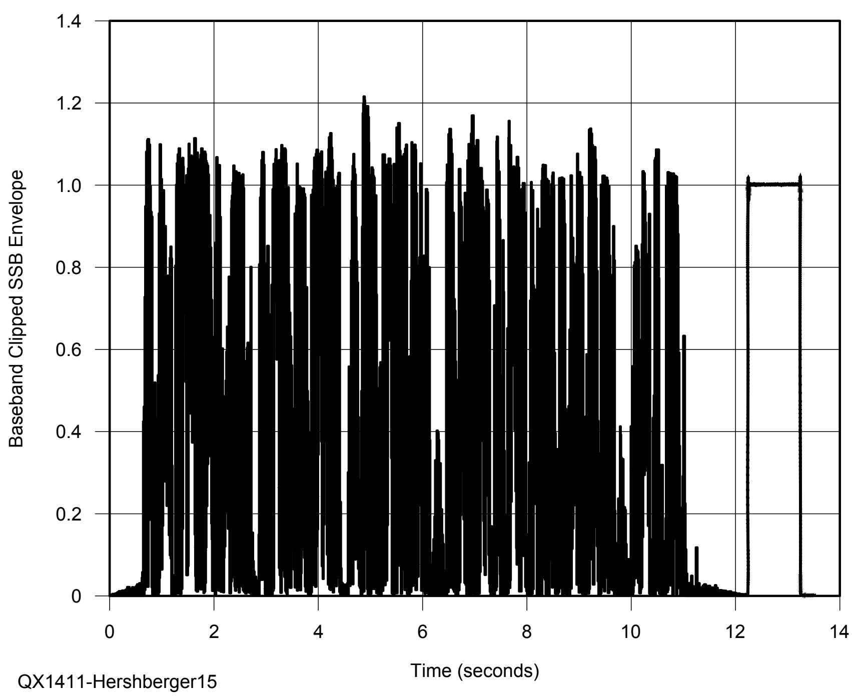


Figure 15 — Envelope of conventional SSB Generator with Baseband RF Clipping. Overshoot is 21%.

enough to produce no overshoots, the system does nothing to the SSB baseband signal. Only when the envelope exceeds that of the 100% tone will there be overshoot control processing.

It is possible to overdrive the algorithm by driving the baseband "RF clipper" into a few dB of clipping beyond what is necessary to remove the Hilbert transform overshoots. This will provide a further increase in average power. I think it is better to do baseband audio processing, however, and let the CESSB system only remove overshoots. Historically, baseband audio processing was not considered particularly effective for increasing average SSB power — because of envelope overshoots. This is no longer true, with the introduction of CESSB. With baseband audio processing, sophisticated multiband compression and clipping is possible, with better results than a simple single-band RF clipper.

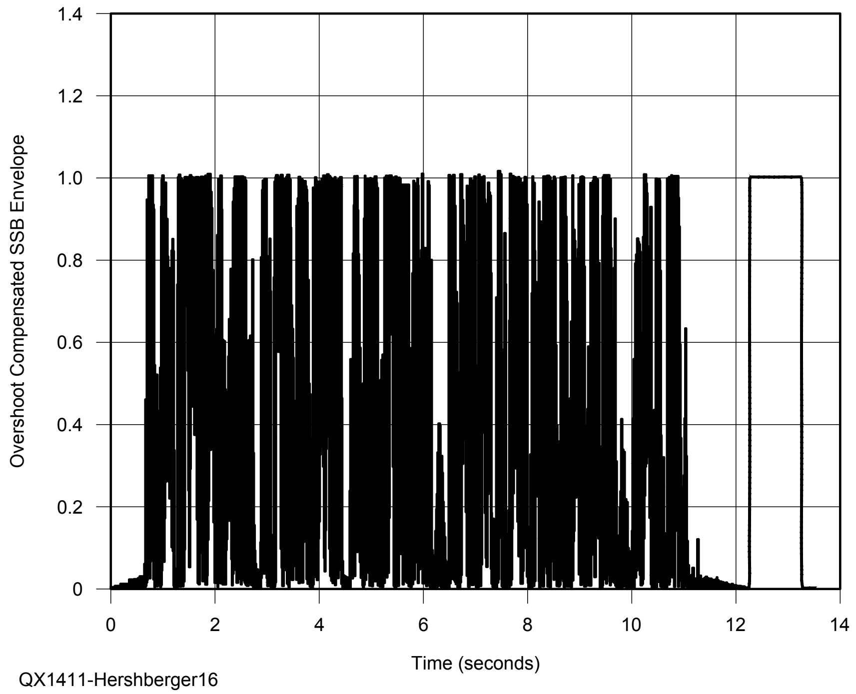


Figure 16 — Envelope of Controlled Envelope SSB Generator. Overshoot is 1.6%

Application Results

FlexRadio made the first hardware implementation of CESSB and reports a 2.56 dB increase in average power over fast look-ahead ALC. Figure 17 shows average power as a function of time for CESSB versus ALC.

CESSB is intended for use with speech signals. Although it is a nonlinear process, the nonlinearity has a negligible effect upon speech. For non-speech SSB applications, however, such as digital modes (PSK, JT65, SSTV, and others), CESSB should be tested to be sure that any average power increase is not offset by nonlinear distortion.

Conclusions

SSB generation generally results in large envelope peaks well above the reference level set by a steady-state tone, even when the input audio is accurately peak limited. SSB envelope overshoot is caused by spectrum truncation and nonlinear phase shifts (particularly from the Hilbert transform). Use of "RF clipping" reduces, but does not eliminate the overshoots. With overshoot compensation, about 3.8 dB of peak reduction is possible using this test speech sample. If power is set to keep peak envelope power the same, this results in an average power increase of about 140%!

Comparisons to a well-designed fast look-ahead ALC system have resulted in an average power increase of about 80% —

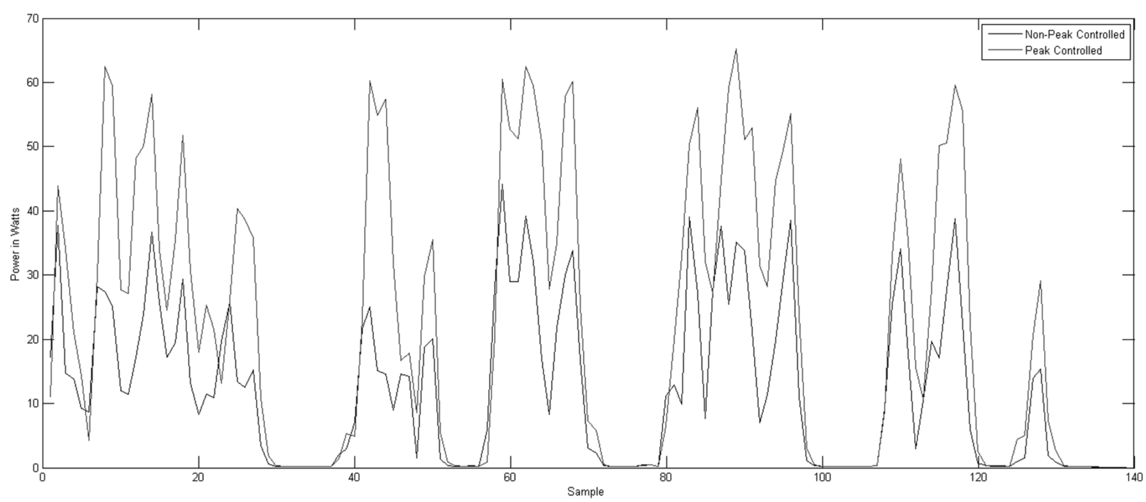


Figure 17 — Average SSB power with fast look-ahead ALC (lower trace) and CESSB (upper trace). (Graphic supplied by FlexRadio)

about 2.56 dB. (That is about the same as converting a single beam antenna to a stacked array.) Your wattmeter should read the same PEP as before, but the average power will be higher.

The processing used does not produce any significant audible artifacts. It may be used in conjunction with speech processing.

The final step in the Controlled Envelope SSB process is a sharp band-pass filter — guaranteeing bandwidth limiting in addition to accurate amplitude control.

In summary, Controlled Envelope SSB envelope control is accomplished by:

1) Prefiltering and peak limiting the audio input signal.

2) Baseband “RF clipping” of the SSB signal to reduce Hilbert transform overshoots.

3) Overshoot compensating the remaining envelope peaks resulting from baseband “RF clipping.”

By accurately controlling SSB envelope peaks at the point where the SSB is generated, ALC is unnecessary. ALC, even with look-ahead, reduces transmitted power when it does not have to — before and after an envelope peak. Average transmitted power can be significantly higher, without introducing “speech processor” type artifacts. An SSB signal with well-controlled envelope peaks makes more efficient use of the RF power amplifier, and produces higher average power for a given peak envelope power.

Acknowledgment

I would like to thank FlexRadio for reducing CESSB to practice, and for verifying its usefulness both with lab tests and real on-air testing.

Notes

[1]Ward Silver, NØAX, Editor, *The 2014 ARRL Handbook*, ARRL, 2013, Chapter 13, Section 1.4, Figure 16.13B, p 13.6. ISBN: 978-1-06259-001-7; ARRL Publication Order No. 0007, $49.95. ARRL publications are available from your local ARRL dealer or from the ARRL Bookstore. Telephone toll free in the US: 888-277-5289, or call 860-594-0355, fax 860-594-0303; **www.arrl.org/shop**; **pubsales@arrl.org**.

[2]You can find more information about *SOX* and download a copy of the *SOX* software at: **sox.sourceforge.net/**.

[3]There is more information about GNU *Octave* on the *Octave* home page at **www.gnu.org/software/octave**. You can also download the latest version of GNU *Octave* from that website.

[4]The supporting files that accompany this article are available for download from the ARRL *QEX* files website. Go to **www.arrl.org/qexfiles** and look for the file **11x14_Hershberger.zip**.

[5]Theodore A Prosch, DL8PT, “A Minimalist Approximation of the Hilbert Transform,” *QEX*, Sep/Oct 2012, pp 25 – 31.

Dave Hershberger, W9GR, was first licensed in 1965 at age 14 as WN9QCH. He is an ARRL Life Member. Dave holds a bachelor’s degree in math from Goshen College and bachelor’s and master’s degrees in electrical engineering from the University of Illinois. He has been awarded 19 US patents. Dave is Senior Scientist at Continental Electronics. His recent projects include two ATSC digital television broadcast exciters with adaptive linear and nonlinear precorrection, a DSP based FM/HD Radio® exciter with adaptive precorrection, new high power uplink transmitters for the JPL/NASA Deep Space Network, and a 2.4 megawatt VLF transmitter system.

Colin M. Brackney, KR4HO

PO Box 301, Lake City, FL 32056; **kr4ho@hotmail.com**

A Receiving Converter for Two-Meter Radios

Extend the receive coverage of that old 2 meter radio for 18 to 74 MHz and 222 to 225 MHz reception.

A few years ago a fellow ham friend left behind a Kenwood TM-241A 2 meter mobile radio in my workshop after paying me a casual visit. It sat on the shelf untouched until recently, when I decided to see whether it worked, or if I needed to repair it. Well, it didn't work. It had a myriad of problems, mostly in the display, where it had burned out bulbs and the digital display was blank. I took it apart, cleaned the intermittent display connections and replaced some of the bulbs. After a few hours of work I had the radio fully functional and on the air.

I had no real use for this radio since I already have a stack of perfectly working 2 meter FM mobile radios of similar vintage that I keep around "just in case." What my stack of radios does *not* have is AM aircraft receive capability that the TM-241A has. These other radios only tune from 136 to 174 MHz and are FM only. The TM-241A receiver, however, tunes from 118 to 135.995 MHz in AM mode and from 136 to 174 MHz in FM mode. It will automatically switch modes between the two ranges. At this point I thought, "Wouldn't it be cool if I could make this thing tune from 18 to 74 MHz so I could listen to 10 and 6 meters?" So then came the ideas that formed the basis for this article, which also includes provisions for receiving the 222 MHz band.

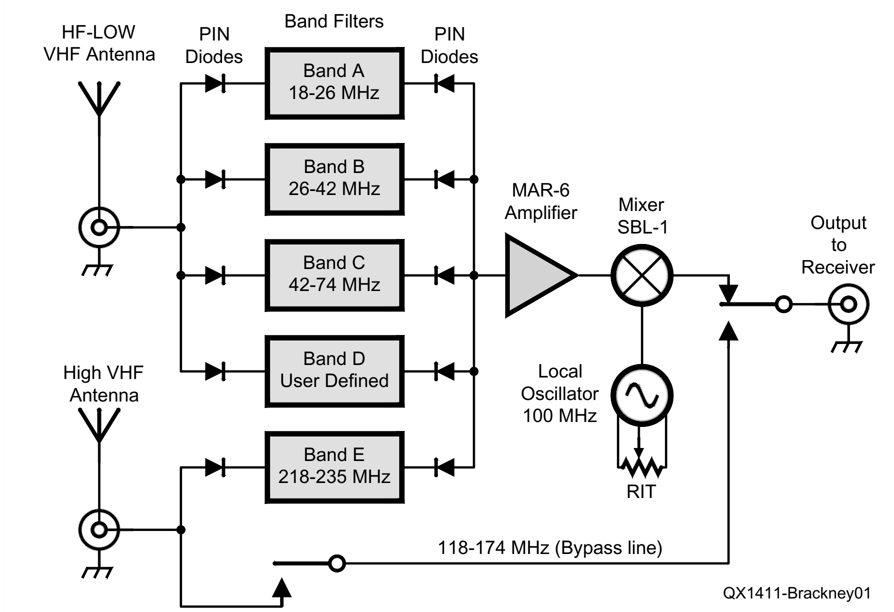


Figure 1 — Here is a simplified block diagram of the receive converter.

Frequency Conversion and the Tunable IF

In earlier times hams often used receive converters in front of a homebrew or surplus 80 meter receiver to allow them to receive 20, 15, and 10 meters. An example would be a tuned stage for 28.0 to 28.5 MHz coupled to a mixer fed with a crystal controlled local oscillator at 24.5 MHz. The resulting output from the mixer would be 3.5 to 4.0 MHz (28.0 MHz – 24.5 MHz = 3.5 MHz). Post-filtering on the converter could filter out most of the 24.5 MHz signal from the local oscillator and would be largely ignored by the 80 meter receiver. Other bands could be received using a single converter and the 80 meter receiver by simply switching in different crystals and corresponding front-end filters or by tuning a preselector.[1] Most of the performance characteristics of the 80 meter receiver were retained when receiving on these other bands.

Multi-band receivers were often built using this technique with the converter modules integrated within the receiver for the desired bands.[2] Some commercially manufactured communications receivers were designed to use optional converters as an accessory for reception of VHF and UHF. Another example would be the satellite or

[1]Notes appear on page 23.

Table 1
Band D Filter Specifications

	T1 Primary/T2 Secondary		T1 Secondary/T2 Primary				
MHz	*Core*	*Link Turns*	*Turns*	*Value*	*C43/44*	*C45/46*	*C38*
21.35	T30-6	2	25	2.3 μH	20 pF	10 pF	1.5 pF
29.35	T30-6	2	24	2.1 μH	20 pF	N/A	1.5 pF
52.0	T30-6	2	13	0.67 μH	20 pF	N/A	1.0 pF
70.0	T30-10	2	12	0.37 μH	20 pF	N/A	1.0 pF

T1 secondary and T2 primary are wound with #32 enameled wire. The links are wound with #26 enameled wire. Toroid cores are available from Amidon.
C38, C45, and C46 are ceramic capacitors.
C43 and C44 are 5-20 pF miniature variable trimmer capacitors. Center frequencies of the filters were calculated using 14 pF for C43 and C44.

microwave receiving system that has the low noise block (LNB) down-converter on or near the antenna. In this situation the feed line carries the much lower frequency and amplified IF signal to the receiver (or tuner), usually located some distance from the antenna.

The converter described in this article takes advantage of the wide tuning range of the 2 meter receiver, in this case using it as a tunable IF. Up-conversion is used to receive the HF and lower VHF frequencies, and down-conversion is used to receive the 222 MHz band. Separate individual converters are not needed here, allowing us to use a single integrated local oscillator. Most of the components in this project are inexpensive and easy to find from the usual nationally known parts vendors. One pricing exception is U5, the 100 MHz local oscillator that, at the time of this writing, cost about $29. The other is U2, the SBL-1 mixer that can cost around $6 to $10 depending on the vendor.

The Circuit

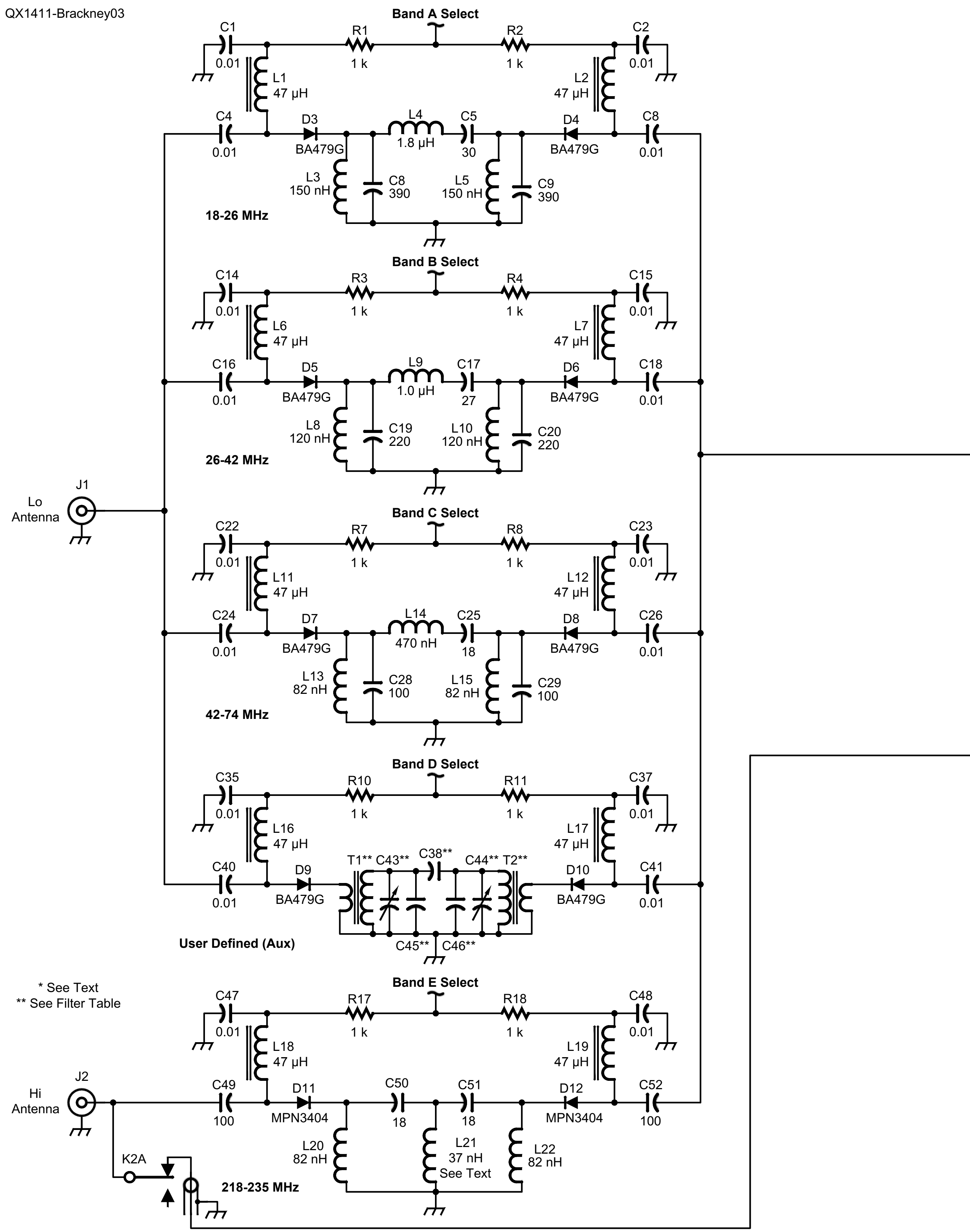
The receiver converter was designed with versatility in mind. You could easily add or delete any of the filter stages to suit the individual listening requirements. A simplified block diagram of the receive converter is shown in Figure 1. The 18 to 74 MHz receive range is divided into three ranges via bands A, B, and C using 3-pole band-pass filters for the specified sub-ranges. I considered using a single low-pass filter for the 18 to 74 MHz range with a roll-off just above 74 MHz. I didn't like the thought of the barrage of strong signals in that wide piece of spectrum all bombarding my front end at once, so I decided to slice up the spectrum a little. The band-pass filters also provide image frequency rejection. I used *ELSIE* to design the filters.[3] These frequency ranges were chosen somewhat arbitrarily. They're supposed have about a 3 dB roll-off at the band edges, so there is a little response overlap. Figure 2 is a photo of the completed converter in use.

The "D" band is an optional *user-defined* band using a double-tuned circuit that is much narrower than the A, B, or C band filters. The double-tuned filter scheme allows the builder to make a simple narrow filter for a specified amateur band of his or her choice. By using the double-tuned narrow filter, receiver overload and intermodulation are reduced greatly by rejecting strong signals from outside of the band. If amateur-band-only reception is desired using this converter, the user could make *all* double tuned narrow filters for each amateur band desired.

Table 1 shows the values needed for 21, 29, 52, and 70 MHz. I did not include values for 18 and 24.9 MHz since AM and FM activity should not found on these bands. For the 218 to 235 MHz "E" band, I eventually opted for a single high-pass filter, as I'll describe later, with a frequency roll-off somewhere below 150 MHz. This also made component selection much easier. PIN diodes D11 and D12 are MPN3404s due to their much lower "ON" state resistance than the BA479G diodes used for the A, B, C, and D bands. The schematic diagram and parts list are shown in Figure 3.

U3, an MAR-6 MMIC, or *Monolithic Microwave Integrated Circuit,* provides the only gain in the receive converter. Its frequency range is specified from DC to 2.0 GHz on the data sheet. The output pin of the MMIC is biased with about 3.6 V, which makes its operating current about 16 mA when R6 is 270 Ω. The gain is stated as being 18 to 20 dB with a noise figure less than 3 dB at the frequencies used in this project. At first I thought this gain figure might be a little excessive until I considered the losses

Figure 2 — This photo shows the receive converter and Kenwood TM-241A radio.


QX1411-Brackney03
Band A Select
C1 0.01
R1 1 k
R2 1 k
C2 0.01
L1 47 µH
L2 47 µH
C4 0.01
D3 BA479G
L4 1.8 µH
C5 30
D4 BA479G
C8 0.01
L3 150 nH
C8 390
L5 150 nH
C9 390
18-26 MHz
Band B Select
C14 0.01
R3 1 k
R4 1 k
C15 0.01
L6 47 µH
L7 47 µH
C16 0.01
D5 BA479G
L9 1.0 µH
C17 27
D6 BA479G
C18 0.01
L8 120 nH
C19 220
L10 120 nH
C20 220
26-42 MHz
J1
Lo Antenna
Band C Select
C22 0.01
R7 1 k
R8 1 k
C23 0.01
L11 47 µH
L12 47 µH
C24 0.01
D7 BA479G
L14 470 nH
C25 18
D8 BA479G
C26 0.01
L13 82 nH
C28 100
L15 82 nH
C29 100
42-74 MHz
Band D Select
C35 0.01
R10 1 k
R11 1 k
C37 0.01
L16 47 µH
L17 47 µH
C40 0.01
D9 BA479G
T1** C43** C38** C44** T2**
D10 BA479G
C41 0.01
C45** C46**
User Defined (Aux)
* See Text
** See Filter Table
Band E Select
C47 0.01
R17 1 k
R18 1 k
C48 0.01
L18 47 µH
L19 47 µH
J2
Hi Antenna
C49 100
D11 MPN3404
C50 18
C51 18
D12 MPN3404
C52 100
L20 82 nH
L21 37 nH See Text
L22 82 nH
K2A
218-235 MHz

QX1411-Brackney03

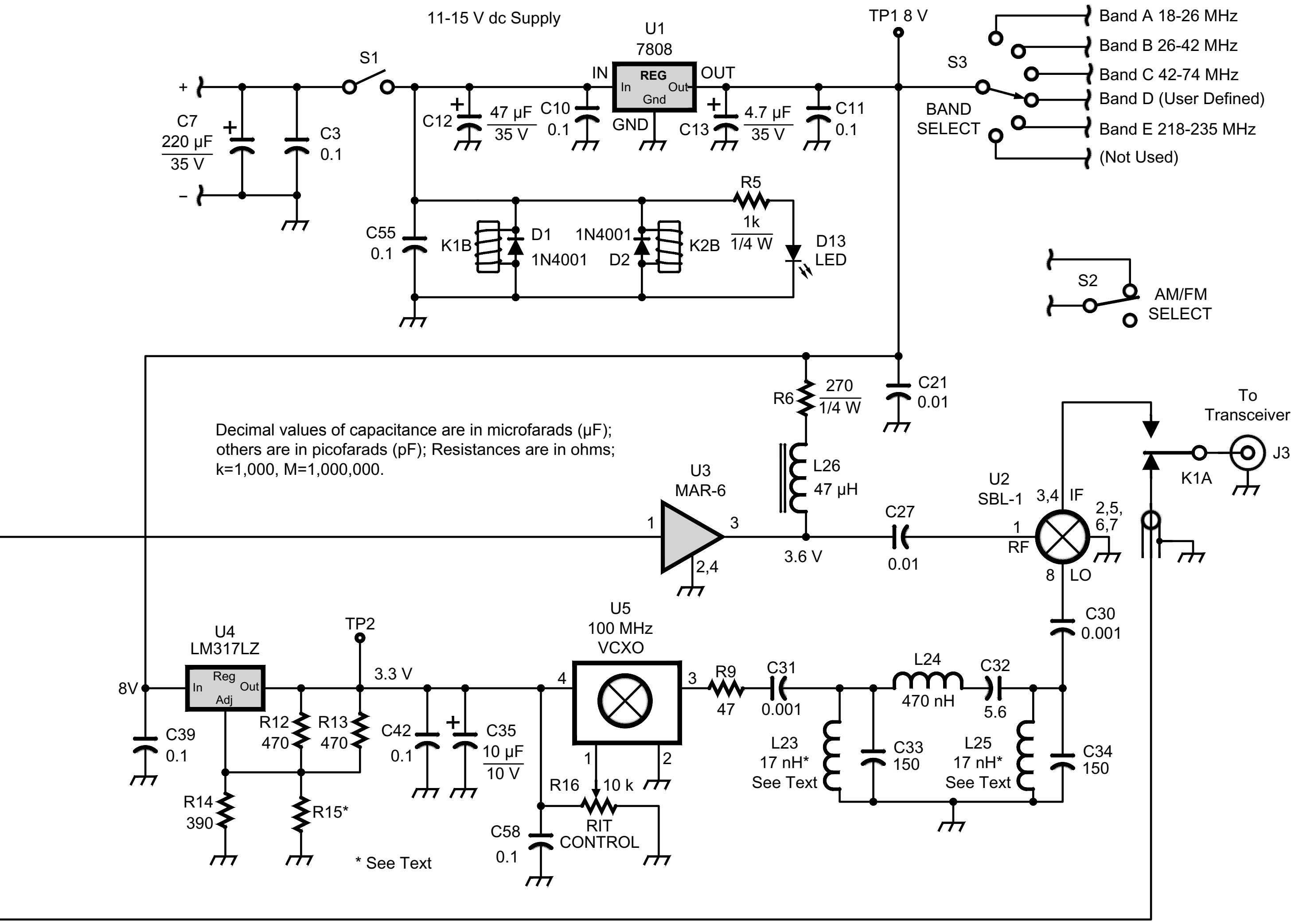


Figure 3 — This is a schematic diagram of the receive converter.

Parts List for The Receive Converter.

Component ID	***Quantity***	***Vendor Part Number***
C1, C2, C3, C4, C6, C14, C15, C16, C18, C21, C22, C23, C24, C26, C27, C36, C37, C40, C41, C47, C48	**21**	**0.01 µF Ceramic Capacitor (Mouser P/N 581-SR215C103KARTR2)**
C38**	**1**	**Ceramic Capacitor – 1.0 pF (Mouser P/N 81-RCE5C1H1R0C0K1H3B) or 1.5 pF (Mouser P/N 810-FK28C0G1H1R5C) See Table 1.**
C45, C46 **	**2**	**10 pF Ceramic Capacitor (Mouser P/N 810-FK28C0G1H100D)**
C43, C44 **	**2**	**20 pF Variable Capacitor (Mouser P/N 659-GKG20015)**
C30, C31	**2**	**0.001 µF Ceramic Capacitor (Mouser P/N 81-RDER71H102K0K103B)**
C10, C11, C39, C42, C55, C56	**6**	**0.1 µF Ceramic Capacitor (Mouser P/N 810-FK18X7R1H104K)**
C35	**1**	**10 µF 25 V Electrolytic Capacitor (Mouser P/N 75-515D106M025JA6AE3)**
C28, C29, C49, C52	**4**	**100 pF Ceramic Capacitor (Mouser P/N 810-FK28C0G1H101J)**
C33, C34	**2**	**150 pF Ceramic Capacitor (Mouser P/N 810-FK18C0G2A151J)**
C25, C50, C51	**3**	**18 pF Ceramic Capacitor (Mouser P/N 810-FK18C0G1H180J)**
C19, C20	**2**	**220 pF Ceramic Capacitor(Mouser P/N 810-FK28C0G2A221J)**
C7	**1**	**220 µF 25V Electrolytic Capacitor (Mouser P/N 667-ECA-1EHG221)**
C17	**1**	**27 pF Ceramic Capacitor (Mouser P/N 81-RPE5CA270J2P1Z03B)**
C5	**1**	**30 pF Ceramic Capacitor (Mouser P/N 150-50N5-300J-RC)**
C8, C9	**2**	**390 pF Ceramic Capacitor (Mouser P/N 810-FK26C0G2J391J)**
C13	**1**	**4.7 µF 25V Electrolytic Capacitor (Mouser P/N 647-UVR1E4R7MDD)**
C12	**1**	**47 µF 25V Electrolytic Capacitor (Mouser P/N 647-UFW1E470MDD)**

Parts list continued on next page.

Component ID	*Quantity*	*Vendor Part Number*
C32	1	5.6 pF Ceramic Capacitor (Mouser P/N 80- C315C569D2G)
D1, D2	2	1N4001 Diode (Mouser P/N 512-4001))
D3, D4, D5, D6, D7, D8, D9, D10	8	BA479G PIN Diode (Mouser P/N 78-BA479G)
D11, D12	2	MPN3404 (RF Parts)
D13	1	LED
L9	1	1.0 μH Inductor (Mouser P/N 871-B78108S1102K)
L4	1	1.8 μH Inductor (Mouser P/N 542-78F1R8-RC)
L8, L10	2	120 nH Inductor (Mouser P/N 807-1025-96KTR)
L3, L5	2	150 nH Inductor (Mouser P/N 434-MICC/N-R15K-RC)
L1, L2, L6, L7, L11, L12, L16, L17, L18, L19, L26	11	47 μH Inductor (Mouser P/N 542-78F470-RC)
L23, L25	2	17 nH Inductor (see text)
L14,	2	0.47 μH Inductor (Mouser P/N 70-IM4-J-.47)
L13, L15, L20, L22	4	82 nH Inductor (Mouser P/N 807-1026-14KTR)
L21	1	37 nH Inductor (see text)
T1, T2**	2	RF Transformer on Toroid core. See Table 1.
**(Coil forms available at AMIDON Associates)		
R1, R2, R3 R4, R7, R8, R10, R11, R17, R18	10	1 kΩ, ⅛ W Resistor (Mouser P/N 299-1K-RC)
R5	1	1 kΩ, ¼ W Resistor (Mouser P/N 291-1K-RC)
R6	1	270 Ω, ¼ W Resistor (Mouser P/N 291-270-RC)
R9	1	47 Ω, ⅛ W Resistor (Mouser P/N 299-47-RC)
R12, R13	2	470 Ω, ⅛ W Resistor (Mouser P/N 299-470-RC)
R14	1	390 Ω, ⅛ W Resistor (Mouser P/N 299-390-RC)
R15*	1	1/8 W Resistor used if U4 output is greater than 3.45 V (see text).
R16	1	10 k Ω Linear Taper Potentiometer (RadioShack P/N 271-1715)
U1	1	7808 (or Equiv.) 3 Terminal 8 V regulator (Mouser P/N 863-MC7808ACTG)
U2	1	Mini Circuits SBL-1 or SBL-1+ RF Mixer (RF Parts, Mini Circuits)
U3	1	MAR-6 or MSA-0686 MMIC Amplifier (RF Parts)
U4	1	LM317LZ, Adjustable 3-terminal Regulator (Mouser P/N 863-LM317LZG)
U5	1	100 MHz VCXO (Mouser P/N 815-ABLNO-V-100-T2)
K1, K2	2	SPDT RELAY, 12 V coil (Radio Shack 275-0241)
SW1	1	SPST Switch (Radio Shack P/N 275-324)
SW2	1	SPDT Switch (Radio Shack P/N 275-325)
SW3	1	6 position switch (Radio Shack 275-034)
SO1, SO2, SO3	3	SO-239 Panel Mounted Coax Jack (Mouser P/N 523-83-1R, RadioShack 278-201)

Capacitors used in the RF filters should be NP0 or C0G types.
Resistors should be 5% tolerance or better.

of the filters, PIN diodes, the SBL-1 mixer and the lower receive sensitivity of the TM-241A outside of the 2 meter band. A simple resistive pi-attenuator can be added between C27 and U2 if the gain is found to be excessive or receiver overload is a problem. I suggest no more than 6 to 10 dB of attenuation.

The signal mixing in the converter is done with U2, an SBL-1 double-balanced mixer. It is a passive device so it has some conversion loss, specified to be less than 6 dB. The nominal impedance of the RF and IF ports is around 50 Ω. It has a frequency range of 1 to 500 MHz. The local oscillator port requires 7 to 10 dBm (5 to 10 mW) for best performance of the mixer. The SBL-1 from Mini-Circuits has been around a long time. I've seen it used in dozens of homebrew projects over the last 3 decades and also in some commercial radios that I have serviced over the years. The SBL-1 is versatile, not only used for frequency conversion, but as frequency multipliers, product detectors and balanced modulators.[4]

For me, the most interesting single component of the converter is U5, the 100 MHz voltage controlled crystal oscillator (VCXO) made by ABRACON. It is available from Mouser at a price of about $29. What is so cool about this oscillator are its excellent phase noise characteristics, frequency stability and the ability to vary the frequency by varying the voltage at pin 1. It is a 3rd overtone crystal oscillator specified for use in military communications and HDTV applications. I found it to be very stable at room temperature even when using the voltage-tuning scheme. I easily justified the $29 price tag when I considered the time and cost of building a comparable stable oscillator with this level of performance. I chose the 100 MHz version simply because it would be easy to read the actual receiver frequency of the radio/converter combination. You subtract 100 MHz from the readout when using bands A-D and add 100 MHz when on band E.

I made two basic assumptions regarding the output impedance and spectral purity of the oscillator. The first assumption is that the impedance is low, say less than 5 Ω, and the second is that the signal would be something resembling a square wave. I reconciled these assumptions by adding a 47 Ω resistor in series with the output, which feeds a 100 MHz band-pass filter. The filter attenuates any harmonics that might be present in the oscillator output. The oscillator-resistor-filter combination yielded almost exactly 7 dBm (5 mW) at the filter output. Perfect.

The typical operating voltage requirements for the oscillator are 3.3 V with a maximum current drain of 35 mA. U4, an LM317LZ voltage regulator, and its associated components provide the power requirements for U5. The 3.3 V from U4 also feeds the hot side of the RIT control R16. The RIT control provides frequency adjustment of about –4.5 kHz to +5.0 kHz from the 100 MHz center frequency, perfectly filling the gaps between the 5 kHz steps of the receiver.

The converter power supply requirements are a DC power source of 11 to 15 V. U1, a 7808 8 V, 1 A regulator provides clean regulated power for U3, the MAR-6 amplifier and U4 the 3.3 V regulator. The 8 V from U1 is also used as the bias voltage for the PIN diodes D3-D10. The 11 to 15 V DC source is used to energize the coils of K1 and K2 and the LED, D13, when the converter is switched on.

If you wanted to listen to a signal, say on 52.525 MHz, you would turn on the converter and the 2 meter radio, and then

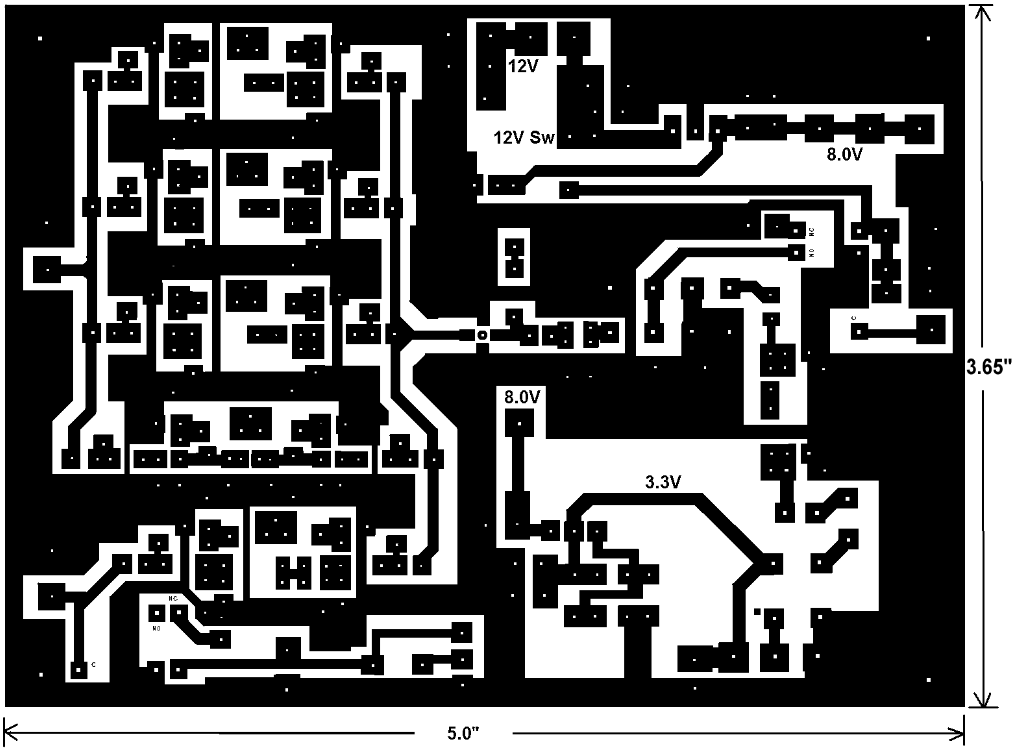


Figure 4 — This is the circuit board etching pattern artwork. The artwork is printed full size for those who want to make their own circuit boards from this pattern.

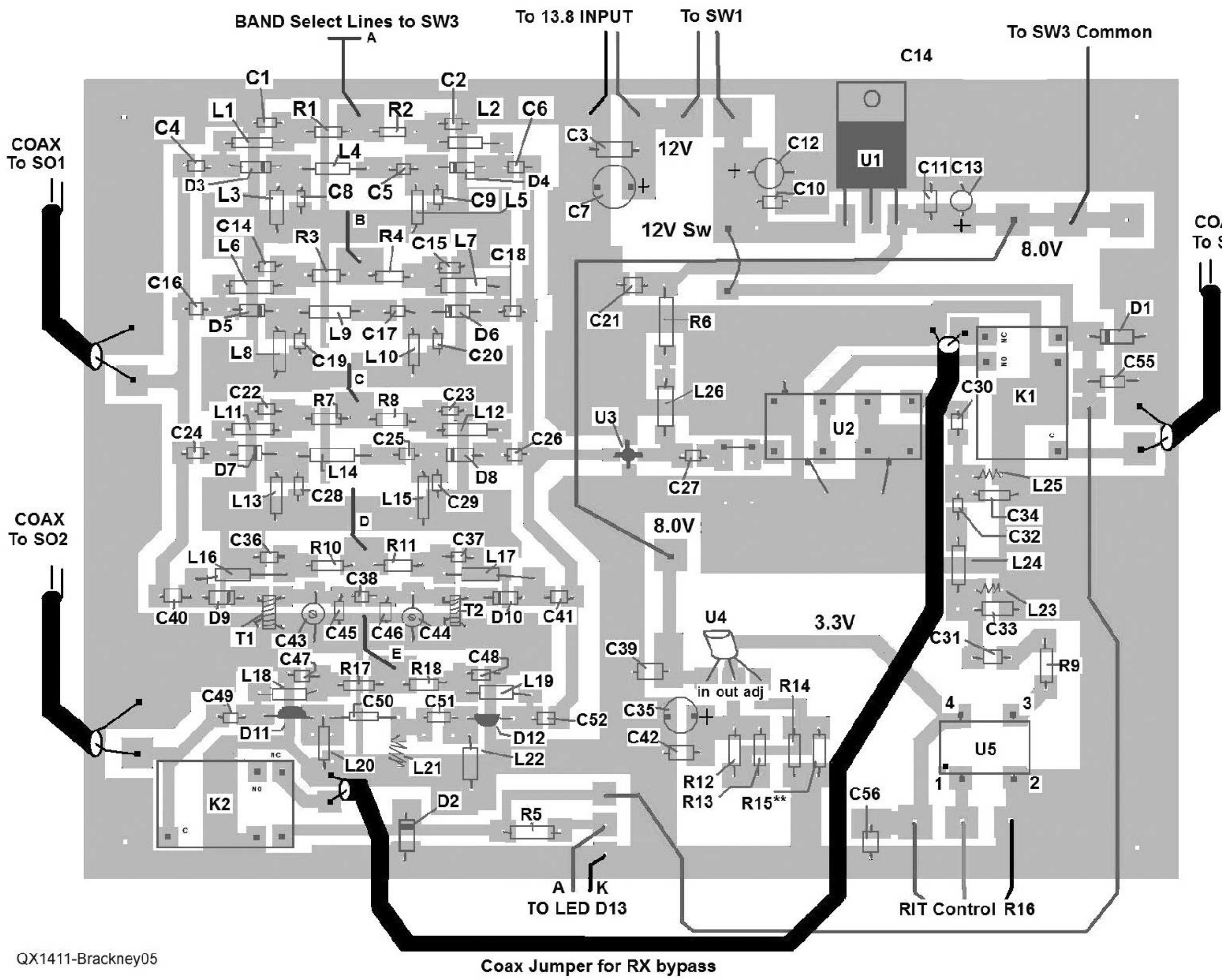


Figure 5 — Here is the circuit board parts placement diagram. Note that the circuit board pattern is shown as an "X-Ray" view from the component side.

you would tune the radio to 152.525 MHz and select the C-range on the converter. The 52.525 MHz signal from the antenna connected to SO1 passes through the filter selected by switch SW3. The voltage applied by SW3 forward biases the PIN diodes D7 and D8 by applying about 7.5 ma through each of these diodes through R7, R8, L11, and L12. The inductors are used to prevent RF loading by the 1 kΩ resistors that provide current to the PIN diodes. The signal passes into and out of the filter, ignoring the other filters since their PIN diodes are not biased and are considered open circuits to the RF.

The RF then passes through DC blocking capacitor C26 to the input pin of U3, the MAR-6 MMIC. R6 limits the operating current of the device to about 16 mA for U3. The amplified RF is applied through DC blocking capacitor C27 to U2, the SBL-1 mixer. The 52.525 MHz signal is mixed with the 100 MHz local oscillator signal from U5 and the 100 MHz band pass filter. The resulting outputs from U2 will occur at the sum and difference frequencies of the RF and the local oscillator. The LO + RF is 152.525 MHz and the LO – RF is 47.475 MHz. The radio will hear the 152.525 MHz (sum) signal and ignore the image (difference) at 47.475 MHz.[5] The IF signal from U2 is fed through the normally open relay contacts of K1 to SO3, which is connected to the radio.

If you wish to receive a signal on 223.500 MHz, you would tune your radio to 123.500 MHz and select the E Band, and the antenna would be connected to SO2. The normally open contacts of K1 pass the signal to the E band filter. The main difference here is that we are tuned to the difference (RF – LO). The sum frequency would occur at 323.500 MHz, which again, our radio would ignore.

When the converter is switched off, K2 and K1 bypass the antenna connected to SO2 to SO3 via a short run of RG-174 coax cable, for normal reception on the 2-meter receiver.

It's important to note here to *NEVER transmit into this converter at ANY power level even when it is turned off.* The relays are not specified for RF use and the circuitry that you worked so hard to assemble will be most likely damaged. Your radio may be a "receive only" or "does not transmit" find from a second hand source or online auction site. I suggest verifying that condition first and disconnecting all leads to the final power amplifier module anyway. Relays K1 and K2 were designed into this circuit to provide RF isolation from the converter when using 2 meters with the power turned off. Last but not least, don't connect the microphone when using the converter.

Building It

I built my converter using a single sided printed circuit board. I suggest using a circuit board or point-to-point wiring for your converter but you may use any method that suits you. Figures 4 and 5 show the printed circuit artwork that I used in this project. Just make sure all of the RF wiring is as short as possible. One of my favorite methods of building circuitry is to use *Vectorboard* prototyping breadboard material with 1/10 inch hole spacing with one side of the board 100% copper clad.[6] The components are mounted on the copper side. The holes for the component leads are cleared of copper by using an oversized drill bit turned by hand to scour the copper out from around the hole to prevent shorts. All component leads that are grounded are simply soldered to the copper surface. This method makes for very short RF ground paths and eliminates the need for a DC ground bus or ground return. All other point-to-point connections are done underneath the board.

When ready to assemble the circuitry, start with the power supply circuits. First the 8 V regulator circuit, then to the 3.3 V regulator, testing each stage as it's completed with an 11 to 15 V input supply voltage. The 3.3 V circuit is of special interest since it feeds the 100 MHz local oscillator. If the output voltage of U4 is greater than 3.45 V then R15 must be installed in the circuit. The value of R15 must be determined experimentally by temporarily connecting a 10 kΩ potentiometer across R14 and adjusting it until U4 has an output of 3.3 V. Carefully disconnect the potentiometer and measure the value with your ohmmeter. For R15, choose a fixed 5% or better tolerance resistor closest to the value of the potentiometer you just measured. After installing R15, check the output of U4 to be sure the voltage is now between 3.2 and 3.45 V. Since U4 powers U5 — that $29 oscillator — it's important to make sure the voltage is right. I did not need R15 since my U4 voltage came to 3.35 V without it.

Next, build the front-end filter sections. Refer to the values in Table 1 for any narrow filters you may wish to use. L21 in the 222 MHz "E" Band filter is a 37 nH inductor made by close winding 5 turns of #32 enameled wire on a 1/8 inch drill bit. After your filter sections are assembled you may wish to test them. I used my MFJ-259B antenna analyzer to test the low-VHF filters. My method was to temporarily solder a 47 Ω resistor across the input pad of U3 and ground. Do this before installing U3.

Manually select each filter by clipping a jumper from TP1 (the 8 V test point) to each of the filter Band Select lines one at a time. Measure the DC voltage across each of the 1 kΩ biasing resistors that feed current to the PIN diodes. The measured voltage should be about 7.5 V ± 0.2 V. Connect a piece of RG58/U or smaller 50 Ω coax with a PL-259 (or N) connector on one end by soldering the cut-end center conductor to the Lo-VHF input and the shield to ground. Feed the antenna analyzer to the PL-259 (or N) connector. Select each of the filters you've built as you did when you tested the diodes. A 3:1 SWR is okay at the band edges but should drop to less than 1.5:1 or better in the middle. You can use the same method to check the 222 MHz band filter if you have an analyzer that can measure this frequency range. I don't, so I waited to test it when I completed the converter. When you're done with this test, remove the temporary 47 Ω resistor.

Build the 100 MHz Local Oscillator Filter using C30-C34 and L23, L24, and L25. Inductors L23 and L25 are hand-wound 17 nH inductors made by close winding 4 turns of #32 enameled wire on a 1/8 inch diameter drill bit. Once completed, this filter will need some adjustment. If you have an SWR analyzer that can measure 100 MHz, connect it like you did when you tested the front-end filters by feeding the SWR analyzer to the open (oscillator) end of C31 and ground. Connect the 47 Ω resistor across the mixer end of C30 and ground. Set the analyzer to 100 MHz. Adjust L23 and L25 alternately with a sharp toothpick by spreading apart (or squeezing together) the turns of the inductors. You should be able to achieve a 1.5:1 SWR or better. If not, try bending C32, C33, and C34 down some, being careful not to break any leads. Once you have achieved your sweet spot with the filter, tack the parts in place with small amounts of epoxy or hot glue.

Leave the 47 Ω resistor in place if you wish to test the oscillator. The oscillator is an SMT device, which makes installing it here a little tricky.[7] If you use the circuit board, solder thin jumper wires (I suggest bare wire-wrap wire) from the SMT pads into the holes on the circuit board. If you use the Vectorboard method, scrape off any copper where the SMT pads would rest except for under pin 2. Solder pin 2 to ground and solder thin wires to the pads on the oscillator making sure they protrude underneath the Vectorboard. Complete the rest of the associated oscillator circuitry. Connection of R16, the RIT control is not needed for this test. If using an RF voltmeter, you will want to see at least 0.45 V to 0.7 V RMS across the 47 Ω resistor when testing the oscillator. I used a calibrated homebrew power meter that showed 5 mW RF power at the filter output.

Special attention should be paid to U3,

the MAR-6 MMIC amplifier. Pin 1 is the input and is the only lead cut at an angle at the factory. The MMIC can be mounted on top or underneath the circuit board. It all depends on your construction method. Mine is mounted right side up, but underneath the circuit board since the leads are short. I drilled a hole in the circuit board for the top of the IC to rest in. If using Vectorboard, you may have to use the same method you used for connecting the oscillator. U3 will be mounted on top of the board, in this case with bus wire connecting the input and output leads. Make sure both leads 2 and 4 are soldered to ground. U2 has pin 1 marked with a different color on the bottom (usually blue). The pins are *not* lined up like they are on a DIP IC. Refer to the datasheet from Mini-Circuits for the pin-out data for the SBL-1 or SBL-1+.[8] Solder the metal case of U2 to ground in 2 or 3 places on the circuit using bare or bus wire.

The completed circuit board, as shown in Figure 6 should be mounted inside a metal enclosure of your choice. I chose to use an external fuse for my converter. The unit draws about 150 mA with a 13.8 V input. The DC connections were made using Anderson Power Pole connectors. All of the RF jumpers and interconnects are made with RG-174 coax cable. You can use any 50 Ω RF connectors you wish for SO1, SO2 and SO3. I used all SO-239s. Figure 7 shows the completed receive converter rear connections. The large bolt in the center is for a ground connection. I forgot to label it. Figure 8 is the top view of the opened converter. Figure 9 shows the completed receive converter in use with the TM-241A radio.

The AM/FM select switch, SW2, is an option you'll want if you make the modification to your radio to manually select the receiving mode. The switch is connected to Anderson Power Pole connectors on the back of the enclosure. I chose a 2-pole 6-position switch for SW3 but you may use a single pole version with as few or many positions as needed. I like extra switch positions in case I want to add stuff later.

What Can We Hear?

Band conditions were excellent the weekend I finished building my receive converter. For my first test I connected a 10 meter dipole and dialed down to 20 MHz to see if I could hear WWV. At 10:00 AM from my northern Florida location I could hear the signal loud and clear with very little noise. The audio quality was excellent. I tuned up to around 21.5 MHz and heard several foreign HF broadcasting stations. On 10 meters I heard several hams on AM phone between 29.0 and 29.1 MHz. Many of them were from Europe. The FM portion of the

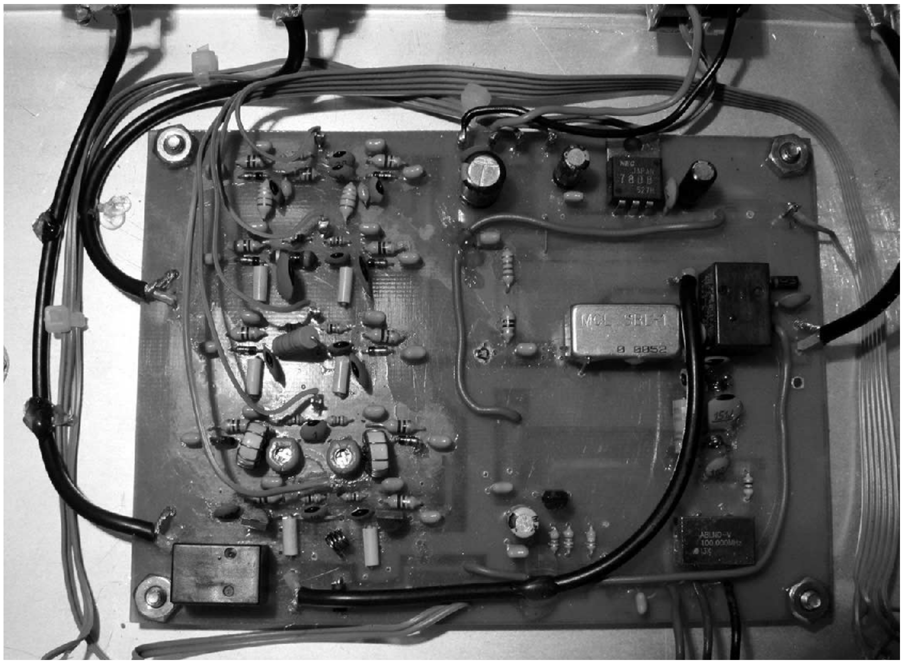


Figure 6 — This photo shows the completed and wired circuit board.

Figure 7 — Here is a photo of the receive converter, viewed from the back. The bolt is for connecting a ground.

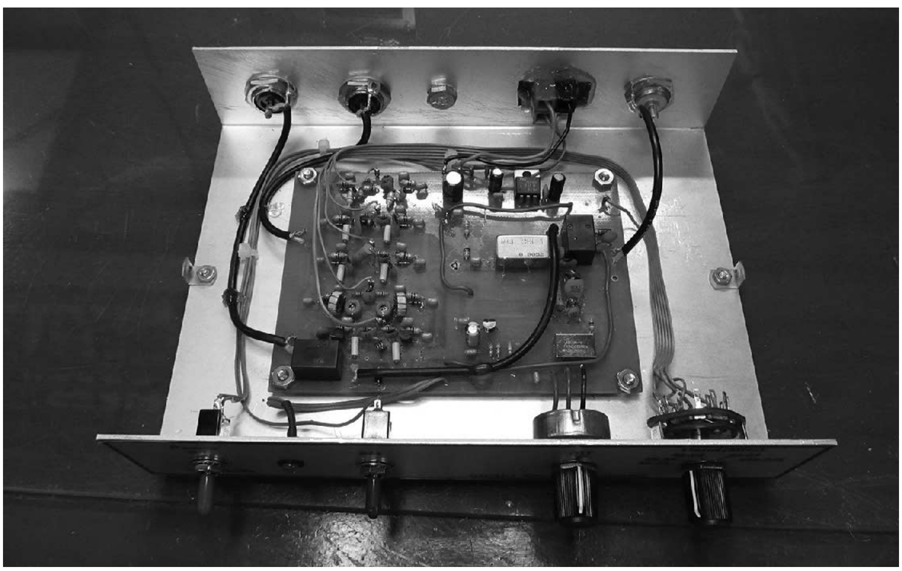


Figure 8 — Here is a top view of the completed receive converter, with the lid off.

band was active and I copied full-quieting simplex and repeater signals from Europe, Canada, and the western US. I also caught a lengthy AM QSO on 15 meters later that same day. The 6 meter band conditions were not so great. It was essentially dead. Using my 6 meter vertical, I tuned to the frequency of the nearest 6 meter repeater and eventually I heard it ID. The signal was as clear as I hear it from my shack radio using the same antenna. I was also able to copy a few brief public-safety transmissions in the 30 to 50 MHz range. There is no known 222 MHz activity in my area and I heard nothing in this range.

Test Bench Realities

On the bench I measured the sensitivity of the receiver and converter combination using the FM mode only. The sensitivity of the TM-241A by itself varied dramatically, depending on where it was tuned. The 12 dB-SINAD FM sensitivity of the receiver at 118 MHz was about 3.0 µV, while it was better than 0.3 µV at 146 MHz. The sensitivity was about 7 µV at 170 MHz. The converter had much better sensitivity on the low VHF range when using these receive ranges, because of the gain provided by U3. At 18 MHz the sensitivity was 0.75 µV, and at 70 MHz it was 2.4 µV. The TM-241A receiver can be aligned to improve these figures but there will be a tradeoff in sensitivity at other frequencies. The 15, 10 and 6 meter ranges all had measured sensitivity from 0.25 µV to 0.35 µV.

When I tested the receiver and converter combination at 223.5 MHz I was disappointed to find the sensitivity at about 25 µV for 12 dB SINAD. Since the Lo-VHF section gave me good and predictable results, I was certain the problem had to be in or around the "E" band filter. I removed the relays, adjusted the inductors in the filter, and tried different filter configurations. What finally made the real difference was to use a high-pass filter with a 3 dB corner frequency at about 110 MHz. The high pass filter was more forgiving with inductor tolerances and made component selection easier. I also changed D11 and D12 to MPN3404 PIN diodes. They have a much lower on state resistance than the BA479G PIN diodes. After those modifications, I measured 0.8 µV at 12 dB SINAD. That is still about 6 dB away from my goal, but I decided it was good enough. One explanation for this could be the loading effect the other filters have on the "E" band filter. The capacitors and diodes that feed the other filters and the geography of the circuit board itself might have an effect. I decided I should be able to hear any reasonably strong signals with a good antenna. Also, I intend to use the converter/receiver in my shop as a monitor receiver.

Final Considerations

The radio I used in this project is an older one. After I repaired it I was able to modify it for manual selection of the AM and FM modes. Some radios don't have an AM mode at all, and some newer ones may allow for manual selection in its user menu. I suggest using a radio that you don't need for anything else, especially if you plan to modify it. You can custom tailor your converter for your specific frequencies of interest.

There are several ways this converter can be built and used. There is a lot of room for any customization that the builder may wish to integrate. Use as many or few filters as you wish or use a different local oscillator for other receive frequency ranges. The circuit is simple enough to easily tweak, and allows for easy experimentation.

Figure 9 — This photo shows the receive converter in action on the author's desk.

AM/FM Manual Select Modification for the Kenwood TM-241A

The details of this modification are offered for those individuals with a TM-241A 2 meter radio who wish to manually select the AM and FM modes using an external switch. I don't have modification details for any other radios at this time. In its original condition this radio will receive the commercial aircraft band from 118 MHz to 135.995 MHz in the AM mode. In the FM mode, it receives Public Safety/Government, Amateur and NOAA Weather frequencies from 136 to 174 MHz. This radio and many other similar models will automatically select the mode for the receive frequency that it is tuned to. The modification will have no effect on the transmitting circuitry of this radio since it only switches between the AM and FM detectors in the receiver. The task is fairly simple, only requiring good eyesight (or a fair amount of magnification) and a steady hand for soldering with a fine tip iron. Go easy on the coffee! I developed this modification with the help of the TM-241A Service Manual.

Be absolutely certain that you want to make this modification. It is reversible, but the one circuit trace you have to cut is thin and will need to be repaired if you want the radio back in its original operating condition. *Make this modification at your own risk.* With all power disconnected from the radio, you need to remove both the top and bottom covers. Disconnect the speaker and set it aside. With the radio front facing you and it sitting right side up, refer to Figure 10, which shows where to cut the trace. In the radio you will notice the same thin circuit trace that comes out from underneath CF1, the 455 kHz ceramic filter. This trace terminates to a thru hole. This area of the main circuit board is located under where the speaker is mounted. *Carefully* cut this trace with a small blade or Exacto knife.

Take a small (½ inch by ½ inch square) piece of circuit board stock. Cut or carve out a line in the copper down the middle, leaving you with two equal copper pads. Next, score one of these copper pads in half. Refer to Figures 11 and 12 for visual details. After verifying that none of these pads is shorted to any of the other pads, solder a 1 kΩ resistor across the two smaller pads. This resistor will limit the current from the 8 V line in the radio in case it gets shorted to ground. Find an open spot on the chassis of the radio and epoxy this board, copper side up, to the chassis. I used the inside wall of the chassis near where the power cabling comes into the radio. This will be the strain relief for the AM/FM switch wires before they exit the radio.

After the epoxy has dried, turn the radio over and refer to Figure 13. The area of interest is located on the opposite side of the circuit board from where you cut the trace. Take two pieces of small wire (I used insulated wire-wrap wire) and cut them long enough to route to the small circuit board you just glued down on the other side. Still referring to Figure 13, solder one of these wires to the *emitter* of Q2 and solder the other to the *base* of Q7. Route these two wires to the small circuit board that you installed and solder the wire from Q2 to one side of the 1 kΩ resistor. Solder the wire from Q7 to the large copper pad. Cut two larger insulated wires (#18 or #20) long enough to lead out of the radio to later connect to the receive converter AM/FM Select switch, SW2. Solder one of these wires to the large pad connected to Q7 and the other to the other side of the 1 kΩ resistor connected to Q2. I found the best way to lead the external switch wires out of the radio is to lift out the power leads where they run through the grommet in the back. Trim some of the rubber from the bottom of the grommet and lead the wires through the vacant space. Check to make sure that none of the connections you just made are shorted to ground before testing the radio.

Once you have verified that you did a good job, reconnect the speaker and apply power to the radio. Turn on the radio and verify that you can hear FM detector noise with the squelch set fully counterclockwise. Short the ends of the two wires leading out of the radio together and the noise should drop much lower, but not go away as the AM detector is switched in. You can now connect these 2 wires to a connector of your choice to interface with SW2 in the converter. I suggest that you tack down all of the new wiring that you just installed in the radio with epoxy or hot glue before putting the radio back together. If you've made it this far and you're smiling, then congratulations! With this modification you can now receive AM or FM at will wherever this radio and the receive converter can receive.

Notes

[1]Lewis G. McCoy, W1ICP, "A Crystal-Controlled Converter for 14 Through 28 Mc," *QST*, Jul 1960, pp 16-18.

[2]Charles Hutchinson, K8CH and Joel Kleinman, N1BKE, Eds, "A High Performance Communications Receiver," *The ARRL Handbook For Radio Amateurs*, 1992 Edition, pp 30.8 – 30.15.

[3]H. Ward Silver, NØAX, Ed, "RF and AF Filters," *The ARRL Handbook For Radio Amateurs*, 2014 Edition, Chapter 11, p 11.9. *Elsie* software is included on the CD ROM that comes with *The ARRL Handbook*. ISBN: 978-1-62595-001-7; ARRL Publication Order No. 0007, $49.95. ARRL publications are available from your local ARRL dealer or from the ARRL Bookstore. Telephone toll free in the US: 888-277-5289, or call 860-594-0355, fax 860-594-0303; **www.arrl.org/shop**; **pubsales@arrl.org**.

[4]H. Ward Silver, NØAX, Ed, "The Double Balanced Mixer: A Basic Building Block," *The ARRL Handbook For Radio Amateurs*, 2014 Edition, pp 10.20 – 10.25.

[5]Joe Carr, *RF Components and Circuits*, 2005 Reprint Edition, Chapter 3, "Radio Receivers: Heterodyning," pp 42 – 44.

[6]*Vectorboard* is available at **www.Mouser.com**.

[7]The datasheet for the ABRACON oscillator is available at **www.mouser.com/ds/2/3/ABLNO-253410.pdf**.

[8]The datasheet for the SBL-1 mixer is available at **www.minicircuits.com**.

Colin Brackney, KR4HO, is an ARRL Member, and has been licensed since 1991. Originally from Yreka, California, Colin developed an interest in radio and electronics when he was 9 years old and began building kits and homebrewing simple receivers. He attended Shasta College in Redding California, where he received his AA degree in Electronics in 1984. Colin began his electronics career in manufacturing at Wiltron Company in Morgan Hill California as a Test Specialist in their Microwave Instruments Division. He later

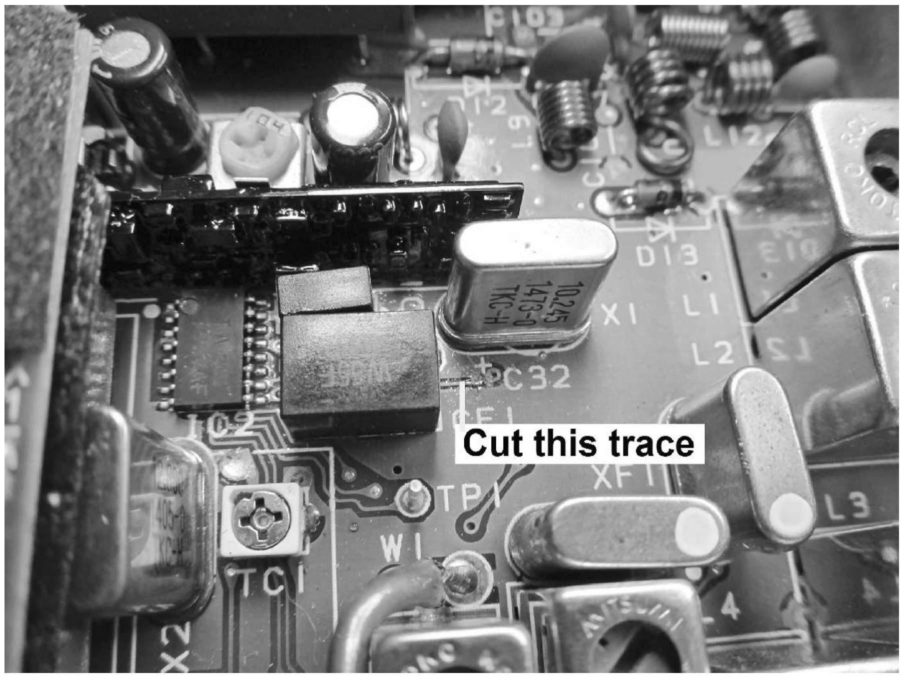


Figure 10 — This photo shows the cut circuit trace inside the modified TM-241A.

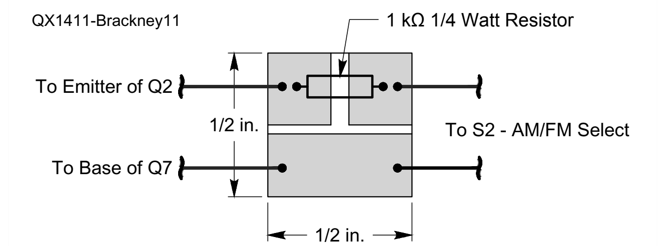


Figure 11 — Here is a pictorial diagram of the circuit addition to the TM-241A.

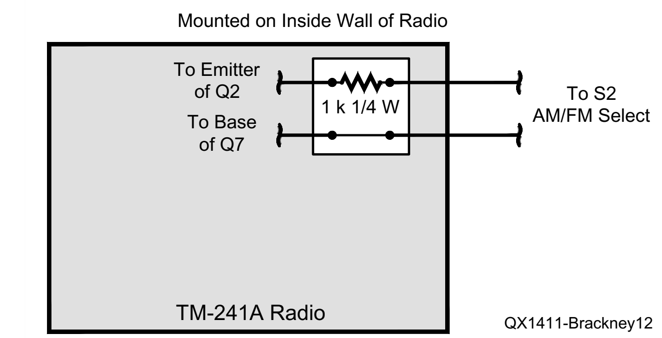


Figure 12 — This is a schematic of the circuit in Figure 11.

worked on microwave small signal amplifiers and YIG oscillators for Avantek in Santa Clara California and then at Ferretec in Milpitas California as a lab technician and later as an engineering technician.

After moving to northern Florida in 1991, Colin took a job as a communications technician for a small commercial two-way radio business. He is currently employed as a telecommunications technician by the Florida Department of Transportation in Lake City Florida, where he has worked on their radio communications systems since 2000. Colin enjoys the hands-on nature of his work, and is an avid homebrewer in his small workshop. His wife Kathleen always knows where to find him. His favorite Amateur Radio activities include ragchewing on 10, 6 and, 2 meters FM and SSB.

The author wishes to thank his wife Kathleen, for her assistance with this article; his friends Randy Pierce, AG4UU and Brian Kopp Ph.D., KC4LPA, for their technical assistance and inspiration; and Jack for the donor radio.

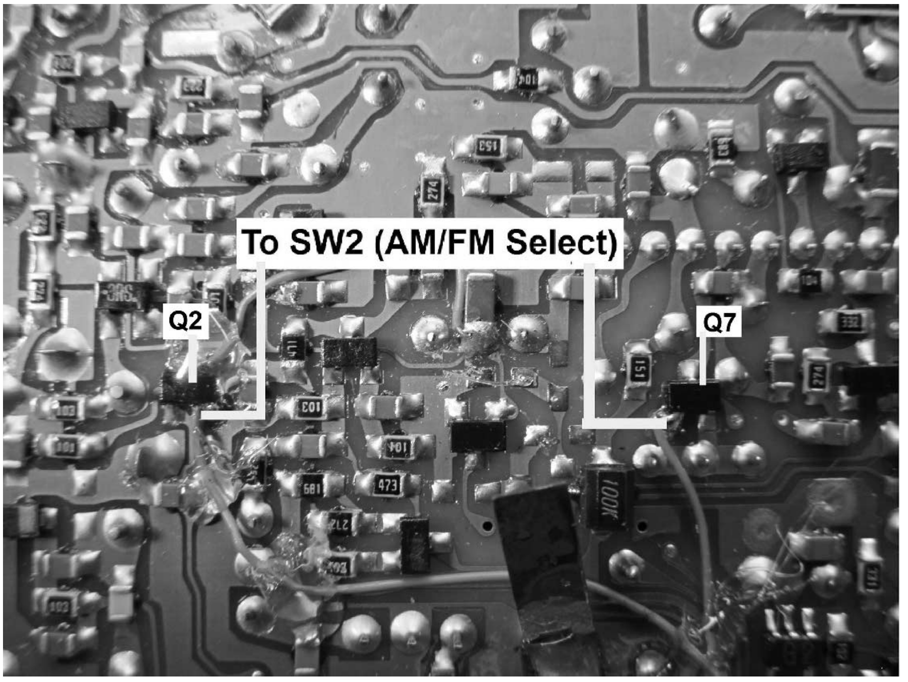


Figure 13 — This photo shows the modified wiring detail on the bottom of the TM-241A main circuit board.

Dr. George R. Steber, WB9LVI

11105 N. Riverland Ct, Mequon, WI 53092; **wb9lvi.arrl.net**

Experiments With Eddy Current Methods for Thickness Measurement of Thin Metallic Materials

Eddy current techniques are often employed to assess the thickness and conductivity of metallic materials. Described here are several techniques for measuring thin metallic layers, such as copper circuit board or aluminum foil, using a simple apparatus or LCR impedance meter.

Recent work in the field of eddy current measurement has sparked the author's interest in the subject, particularly as related to measuring thin metallic plates or copper circuit board material.[1] Over the years, quite a lot of raw copper circuit boards used for projects have accumulated in the workshop and, unfortunately, the copper thickness data is unknown or lost. It would be nice to have a simple way to categorize this material, as it cannot be done easily by visual inspection. Spurred on by the possibility of using simplified methods and low cost measuring circuits, several experiments were undertaken to investigate eddy current methods for thickness measurements. In the course of the work, several simple eddy current sensors and techniques were devised. This includes the design of a phase circuit for thickness measurement and a simple method using an inductance/capacitance/resistance (LCR) impedance meter. Before we proceed too far, however, it is advisable that some background be given in the field of eddy current measurements.

At the heart of an eddy current measuring

[1]Notes appear on page 30.

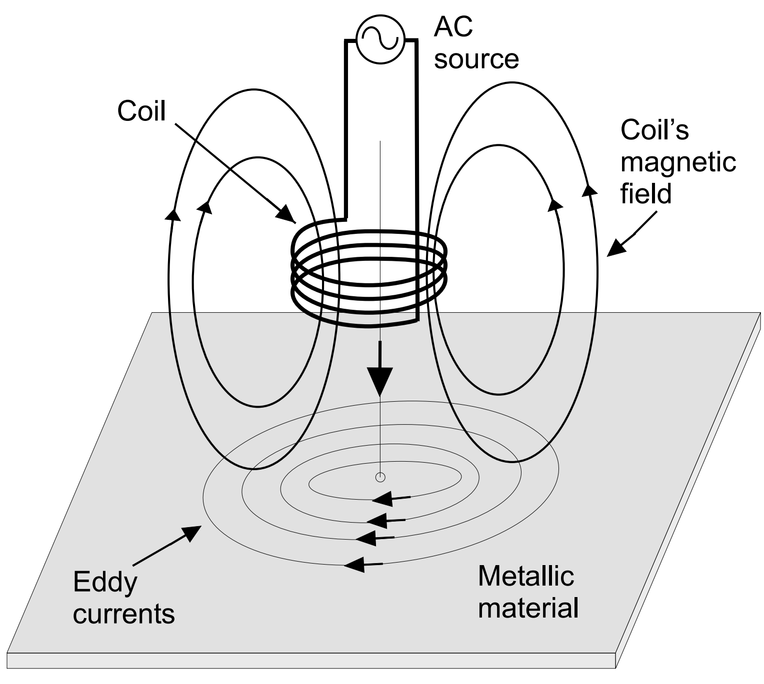


Figure 1 — Shown here is a magnetizing coil producing eddy currents in metallic material.

device is a magnetizing coil with an alternating current in its winding. When the coil is placed near a metallic object, the variable field induces a distribution of eddy currents into the object. Figure 1 shows a simple depiction of such an eddy current arrangement. The amplitude and phase of these induced currents depends on many parameters such as object material, geometry, magnetizing coil, distance to object, exciting signal amplitude and frequency. In turn, these eddy currents generate their own magnetic field that depends on the object geometry and its electrical and magnetic properties.

Stated generally, the resulting eddy current magnetic field contains, in principle, all the characteristics of the object. Determining the object properties from this field is, however, mathematically difficult and has been solved only for simple geometries. Even for these cases, the solutions are complicated and numerical simulation methods must often be employed. An excellent article by Lukas Heinzle on the Internet provides a general introduction to the subject including a complete mathematical treatment.[2] In spite of these complexities, eddy current testing has evolved into a well-developed technology for inspection of thin metals, with many applications in the aerospace, manufacturing, and service sectors.

This article describes the specific application of eddy currents for determining the thickness of thin copper or aluminum foils. It should be mentioned, however, that other methods exist for accomplishing this goal. For example, expensive four-point micro-resistance measuring equipment such as the CM95 or CMI165 can be used for the job in many cases.[3] Our focus here will be mainly on eddy current methods using impedance techniques.

Considerable precision for this task can usually be obtained by using complex instrumentation. The simplified procedure is as follows. First, the impedance of a magnetizing coil in free space, with an AC current in its windings, is measured and saved. Next, the coil is placed on the metallic plate of interest and the impedance is measured again. The difference between the two impedance measurements contains considerable information about the plate thickness and other factors. Of significant interest are the real and imaginary parts of the impedance.

In this article we will take a look at the characteristics of these impedance measurements and discuss simplified ways of measuring thin metallic layers. Particular emphasis will be placed on determining the copper thickness on single-sided copper circuit board material. Suggestions will be made for making your own eddy current sensors by winding your own coils or using inexpensive off-the-shelf inductors. An apparatus for measuring copper thickness will be described as well as the application of an LCR impedance instrument to the task.

Considering the Eddy Current Method

At first glance it may not appear possible that simply putting a coil above a metallic plate as shown in Figure 1 can provide thickness information. Yet, eddy current techniques are used, for example, to measure the thickness of a hot sheet in a rolling mill and metal thinning of an aircraft fuselage due to corrosion. Many different types of sensors are used in these applications. The impedance plane (R versus jX) is often used to display thickness variations and subsurface defects.

Other ways of looking at the impedance data have been considered. Recently Pinotti and Puppin have proposed a major simplification for measuring thickness by concentrating on the phase only. (See Note 1.) They have devised a simple lock-in method using inexpensive CMOS integrated circuits. In order to accomplish this goal they introduced a pickup coil in addition to the field coil. The arrangement is shown in Figure 2. A sine wave signal generator provides the excitation. Amplifier A1 provides the reference signal proportional to the current in the field coil. Amplifier A2 is a high gain amplifier providing the pickup coil signal.

The use of a pickup coil enables more direct measurement of the eddy current field and permits measurement of the phase difference between the free space field (reference field) and the field with the metallic plate. An interesting result is achieved by this arrangement. As the frequency is swept, there will be a peak in the phase difference field. The location of this peak phase, in frequency, corresponds to the thickness of the plate. Results obtained by this arrangement reportedly produced peak phase differences on the order of 80°. So, it should be fairly easy to determine metal thickness by looking for the phase peak versus frequency scan.

Although this method is appealing and apparently works well, the prospect of constructing a two-coil probe dampened my enthusiasm for it. Could similar results be obtained with a single coil? Spurred on by this possibility, I decided to experiment with a single coil configuration and see if there was another way to get the thickness information.

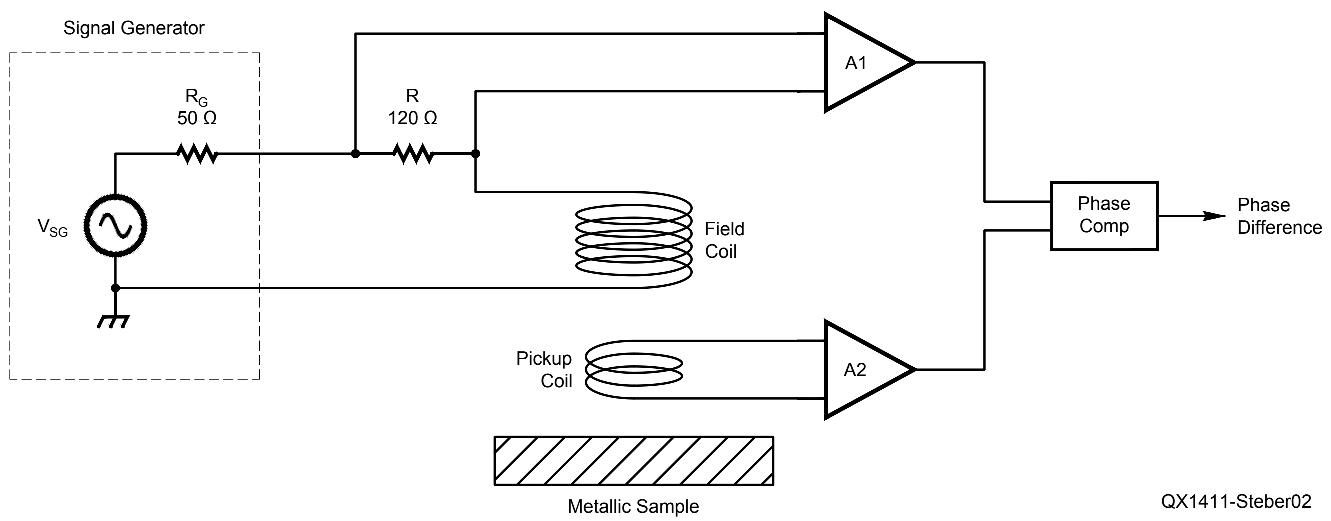


Figure 2 — This is a two-coil eddy-current phase measuring system.

The concept of phase signature, introduced by Yin and others, has been shown to be very useful when assessing plate thickness, as it is very sensitive to thickness and relatively insensitive to other factors.[4] The impedance needs to be measured with good precision, however, and over the proper frequency range. Unfortunately, such measuring devices are usually complex and expensive. Fortunately, in this case, I had constructed a vector impedance analyzer (VIA), similar to the methods described in my LMS Impedance Bridge article in *QEX*.[5] It has proved to be invaluable in this work. Phase signatures obtained with the VIA are useful in seeing how the phase changes with plate thickness.

Obtaining a Phase Signature

Using the VIA makes it is very easy to obtain phase signatures. The basic setup is shown in Figure 3. An important component is the eddy current sensor. I decided to experiment with air-core coils, ferrite-core coils and off-the-shelf inductors. Several sensors are shown in Figure 4. A commercially made inductor is on the left; and air core inductor is in the center; and a ferrite core inductor is on the right. Although the coils are shown on their sides for clarity, during operation, the flat, circular end should make contact with the metallic material being studied.

Air-core coils were the easiest to make. I used old plastic thread spools, about 1 inch diameter for the wire forms. Coils of 80 and 120 turns were fabricated with inductances of 81 µH and 156 µH respectively.

I had several ferrite cores with diameters of 8 mm and 10 mm and approximately 1.5 to 3 inches long on hand. I made many different coils using 30 to 80 turns and having inductances of 40 to over 150 µH.

Finally, I purchased several commercial inductors made by Murata. They were chosen because of their size, low cost, ferrite cores and variety, and because they are available from Mouser. One of these inductors, the Murata 150 µH, 19R154C worked well.

Copper samples in the thickness range expected for single sided copper PCB material are needed for testing. Sheets, or foils, of 1.0, 1.25, 1.4 and 3.0 mils were obtained. The 1.0 mil and 1.4 mil foils were stacked to obtain a 2.4 mil test sample. Each sample is at least 3 × 3 inches. Copper circuit board material is often specified by copper weight. Table 1 shows the relation between copper circuit board material weight and copper thickness in mils and mm for several popular single-sided circuit boards.

I also cut some common kitchen type aluminum foil into small 3 × 3 inch rectangular sheets. According to the manufacturer each sheet is 0.9799 mil (23.62 µm) thick. Foil sheets can be stacked to obtain various thicknesses for test purposes. Since copper and aluminum have different conductivity, the results for one do not correspond to the other. Nevertheless, it is useful to have aluminum material for testing.

Phase Signature For Copper

Quite a lot of tests were run with various sensors. Initial tests with the coils showed that the air-core coils were not as sensitive to phase change as the ferrite coils or Murata inductors. Hence extensive tests with air-core coils were not continued.

Figure 5 shows phase signatures for copper foils obtained with a home brewed eddy sensor. The sensor consisted of a 10 mm diameter by 38 mm long ferrite rod wound with 40 turns of #28 AWG wire. Its inductance was 70.1 µH and resistance was 0.31 Ω measured at 5 kHz. With no copper present (0 mil curve) the phase tends toward 90° as expected for an inductor. There is an interesting region around 5 kHz. All of the curves in this region are separated nicely in phase. That is, there is a phase difference between the copper foils without the curves crossing. In the region from the 1 mil foil to the 3 mil foil there is a phase shift of about 4°. Hence, a sensitive phase measuring device set to 5 kHz should be able to distinguish between the foils based on phase alone.

Figure 6 shows the phase signatures obtained with the 150 µH Murata inductor.

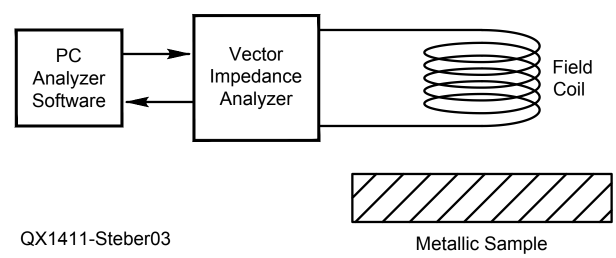


Figure 3 — This diagram represents how I obtained a phase signature using a vector impedance analyzer.

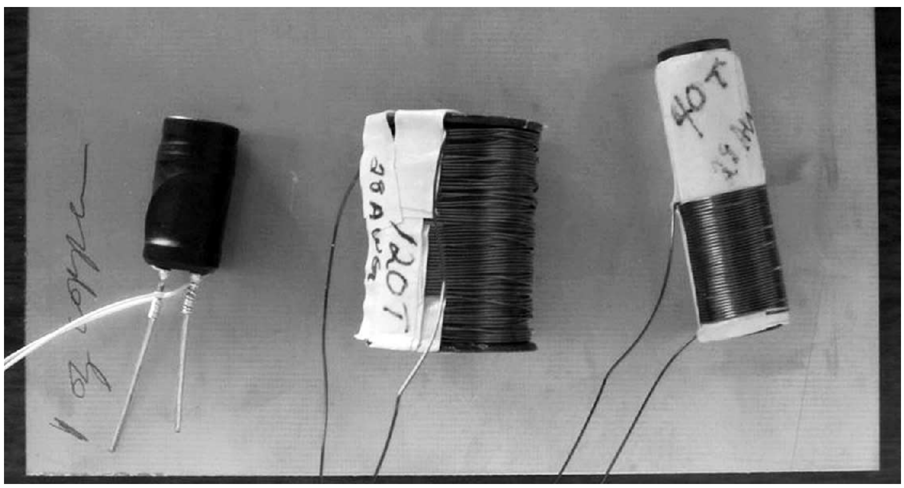


Figure 4 — This photo shows three eddy current sensors. From the left, there is a Murata inductor, an air-core coil, and a ferrite-core coil.

Table 1
Thickness of Some Common Copper circuit board Material.

Circuit Board	*Mil*	*Inch*	*mm*
0.5 oz	0.7	0.0007	0.01778
1 oz	1.4	0.0014	0.03556
2 oz	2.8	0.0028	0.07112

The curves are similar to the previous eddy sensor, but do not have as much phase shift, only about 3.1°. Clearly the design of the sensor is important.

Simple Phase Meter

Figure 7 shows the design of a simple phase meter. It requires the use of a good, audio sine wave generator and a volt-ohm-meter (VOM). The sine wave generator should be able to produce several volts. Connected to the sine wave generator is an isolation transformer. Its function is to provide isolation from the sine wave generator and to provide a common ground reference for the field coil (eddy sensor) voltage and field current. The voltage across resistor R1 corresponds to the current in the coil and provides the reference phase. The field coil voltage provides phase information of the coil. Both voltages are sent to respective comparators U1A and U1B, the outputs of which now contain only phase information.

A 74HCT86 IC (U2A) is used as a phase comparator. Its average phase sensitivity is roughly equal to the supply voltage divided by 180. In this case it is about 28 mV per degree. A balancing potentiometer is used to null the VOM when the coil is in free space. Then the VOM directly reads the change corresponding to the foil thickness. A low-cost, digital readout VOM works well in this application.

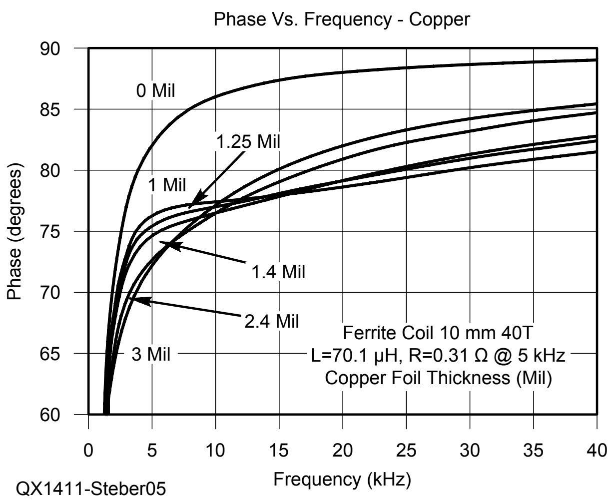


Figure 5 — This graph shows the phase versus frequency plot of a 40 turn ferrite coil on copper.

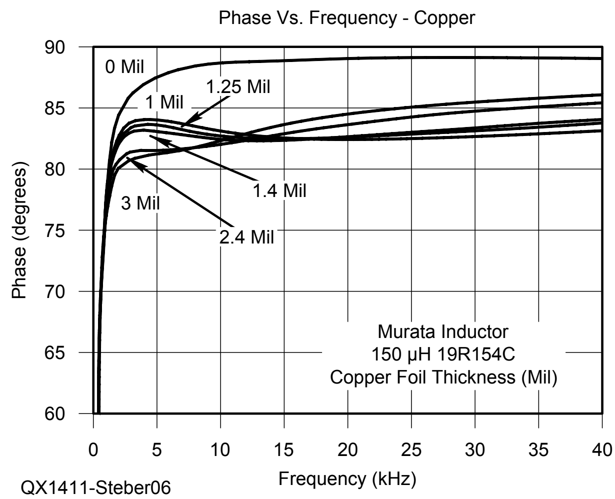


Figure 6 — Here is the phase versus frequency plot of the 150 µH Murata inductor on copper.

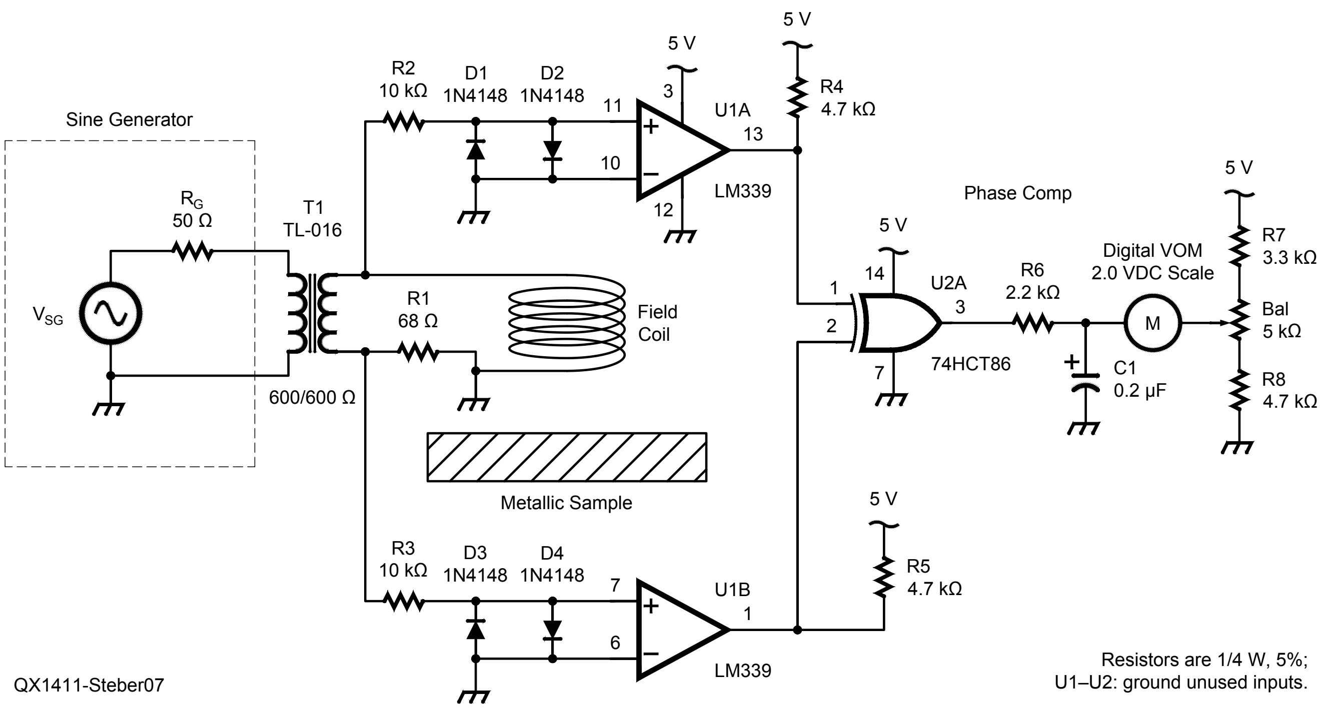


Figure 7 — This schematic diagram shows my phase measuring meter. Transformer T1 is available from Mouser Electronics.

Calibration of the meter is required, using known foil thickness, to correlate the VOM voltages to foil thickness. Thickness versus phase voltage is shown in Figure 8 for two sensors. Once the calibration is complete, the curves may be used to measure unknown copper foil or circuit board thickness.

In practice, this circuit works reasonable well, but there are some things to consider. The circuit requires at least 10 minutes to warm up to minimize drift of components with temperature. A sine wave generator with good amplitude and frequency stability should be used. The supply voltage is important. Since it is used for balancing the VOM, it must be regulated, noise-free and should not drift. In this circuit the coil voltage is small (typically 60 mV pp). Adding an amplifier before the voltage comparator would be an improvement. Using the circuit as shown, consistent readings to within 5 mV were achieved after warm-up.

An important thing to be considered when using an eddy current device is the surface area being measured. It must be large enough to encompass most of the eddy current field lines. Typically at least 1.5 inches must be allowed around the sensor. Also important, the sensor must be in contact with the surface to reduce lift-off effects. As the sensor is moved over the copper, changes in the field will occur due to scratches or other defects in the material. So, you may want to average several readings. And if you get close to the edge of the foil the field will drop off drastically. Nevertheless, if used carefully, this simple circuit can be used to differentiate most single sided copper circuit board material.

Inductance Measurements

Can the inductance variation of the sensor be used to measure copper foil thickness? Figure 9 shows the curves of inductance versus frequency, obtained using the vector impedance analyzer, for the Murata 150 µH inductor. Notice that the region around 10 kHz has a large spread in inductance values, about 14 µH, from 1 mil to 3 mil thickness. If you use a homemade ferrite coil, it will provide an even greater inductance change. Figure 10 shows the calibration curve of inductance versus thickness for a 10 kHz excitation frequency.

Many vector LCR meters can measure inductance accurately at 10 kHz. I have a Tonghui TH2811D LCR meter, which I used to verify this measurement. Its readings compared very well to the vector impedance analyzer data. Many other vector LCR meters are available that work well at 10 kHz. So this appears to be a viable way to measure copper foils.

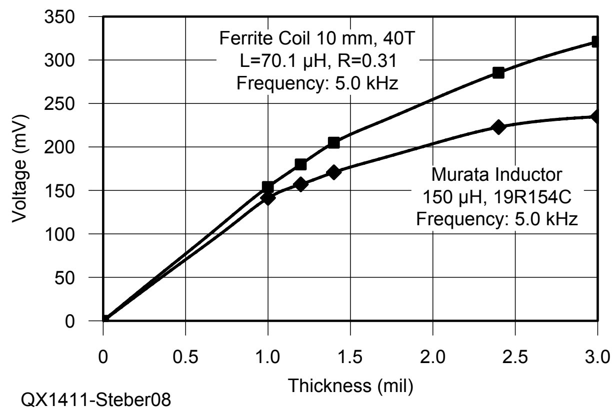


Figure 8 — This graph is the phase voltage versus thickness plot for two sensors on copper.

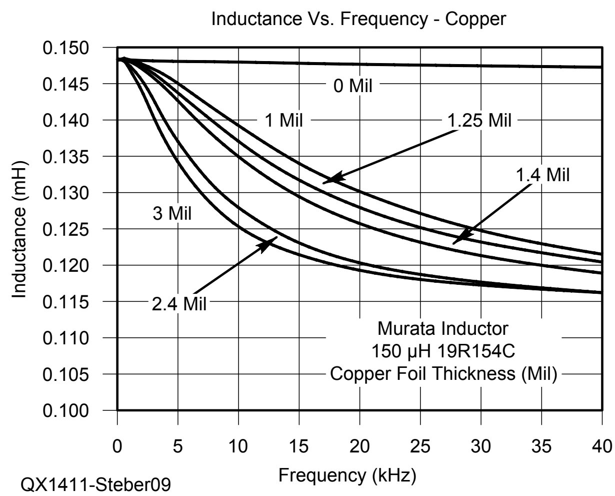


Figure 9 — Here is the graph of inductance versus frequency for the 150 µH Murata inductor on copper.

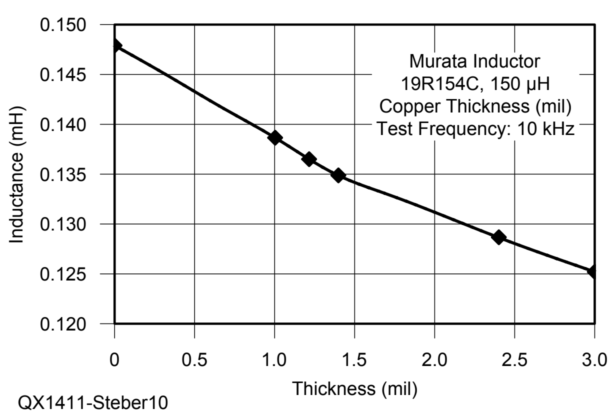


Figure 10 — This graph is the calibration curve of inductance versus thickness for a 10 kHz excitation frequency with the 150 µH Murata inductor on copper.

Summary And Conclusions

This article shows several experiments using eddy current methods for measuring the thickness of copper foils and single sided copper circuit board material. Phase information from the impedance data could be used to differentiate thickness. The article describes a simple phase meter that can be used for this purpose.

The article describes several eddy current sensors. The air-core coil was the least sensitive to phase. The homemade ferrite coils were the best. More research could be done to improve the phase response of the sensors.

Inductance measurements on a Murata 150 µH inductor showed that it could be used to determine copper thickness with proper calibration. Although not as sensitive as a homemade ferrite coil, it can easily be used for this purpose. For those in a hurry, the inductance measurement method will probably be the best, because it only requires an off-the-shelf inductor and a vector LCR meter.

I want to emphasize that this work applies only to single-sided copper circuit board material. Further work needs to be done to apply these techniques for double-sided material. It would be convenient to be able to do the measurement from one side. Since the main magnetic field penetrates deeply for low frequencies, this goal is probably achievable. The calibration method must be modified, however, and many samples of 2-sided circuit board material will be needed. That work will be left to someone down the line. Hopefully positive results will be published in *QEX* at some future time.

Experimenting with eddy currents is enjoyable and educational. In this case they helped me sort out much of my surplus copper circuit board material. Hopefully you have learned something about eddy currents from this project. As you explore the subject further you will find they are useful in a variety of applications. In any event, set your *vector* in the right *field* and let the *eddy current* take you to your goal.

George R. Steber, Ph.D., is Emeritus Professor of Electrical Engineering and Computer Science at the University Of Wisconsin-Milwaukee. He is now semi-retired, having served over 35 years. George, WB9LVI, has an advanced class license, is a Life Member of ARRL and IEEE, and is a professional engineer. His last article for QEX *was "An Unusual Vector Network Analyzer" in the September 2007 issue. George has worked for NASA and the USAF and keeps busy working on various projects at the University. He is currently involved in cosmic ray research and hopes to launch a new program to study them on a global basis. In his spare time he enjoys WSPR/JT9 amateur radio, racquetball, astronomy, and jazz. You may reach him at wb9lvi@arrl.net with "Eddy" in subject line and e-mail mode set to text.*

Notes

[1]E. Pinotti and E. Puppin, "Simple Lock-In Technique for Thickness Measurement of Metallic Plates", *IEEE Transactions on Instrumentation and Measurement*, Feb 2014, Vol 63, No. 2, pp 479 – 484.

[2]Lukas Heinzle, "Eddy Current Testing," Omicron Lab Application Note: **www.omicron-lab.com/bode-100/application-notes-know-how/articles-use-cases/eddy-current-testing.html**

[3]You can find more information about Oxford Instruments coating thickness measuring equipment at: **www.oxford-instruments.com**.

[4]W. Yin and A. Peyton, "Thickness Measurement of Metallic Plates With an Electromagnetic Sensor Using Phase Signature Analysis", *IEEE Transactions on Instrumentation and Measurement*, Aug 2008, Vol 57, No. 5, pp 1803 – 1807.

[5]George Steber, WB9LVI, "An LMS Impedance Bridge", *QEX*, No. 232, Sept/Oct 2005, pp 41 – 47.

Tom Wheeler, NØGSG

10724 Horton, Overland Park, KS 66211; **tom.n0gsg@gmail.com**

New Life for the Motorola MSR-2000 VHF Repeater: A New RF Power Amplifier

Add an RF power amplifier to your Motorola MSR-2000 repeater.

The Motorola MSR-2000 repeaters are popular in both VHF and UHF Amateur Radio applications, and they provide a good balance of performance and ruggedness. Unfortunately, these units are now in their third decade of life, and original replacement parts are hard to come by. In particular, the power amplifiers in these units are typically near the end of their useful lives; they've already given a full life of service as commercial and public safety workhorses, and Amateur Radio is supposed to be a retirement gig for them. Many of the units are just no longer up for the kind of stress we put on them, particularly in hot environments. Thirty years of hard living will do that!

Motorola constructed the PA modules in the VHF systems using proprietary PNP RF power transistors. The amplifiers are of conservative design, and use four

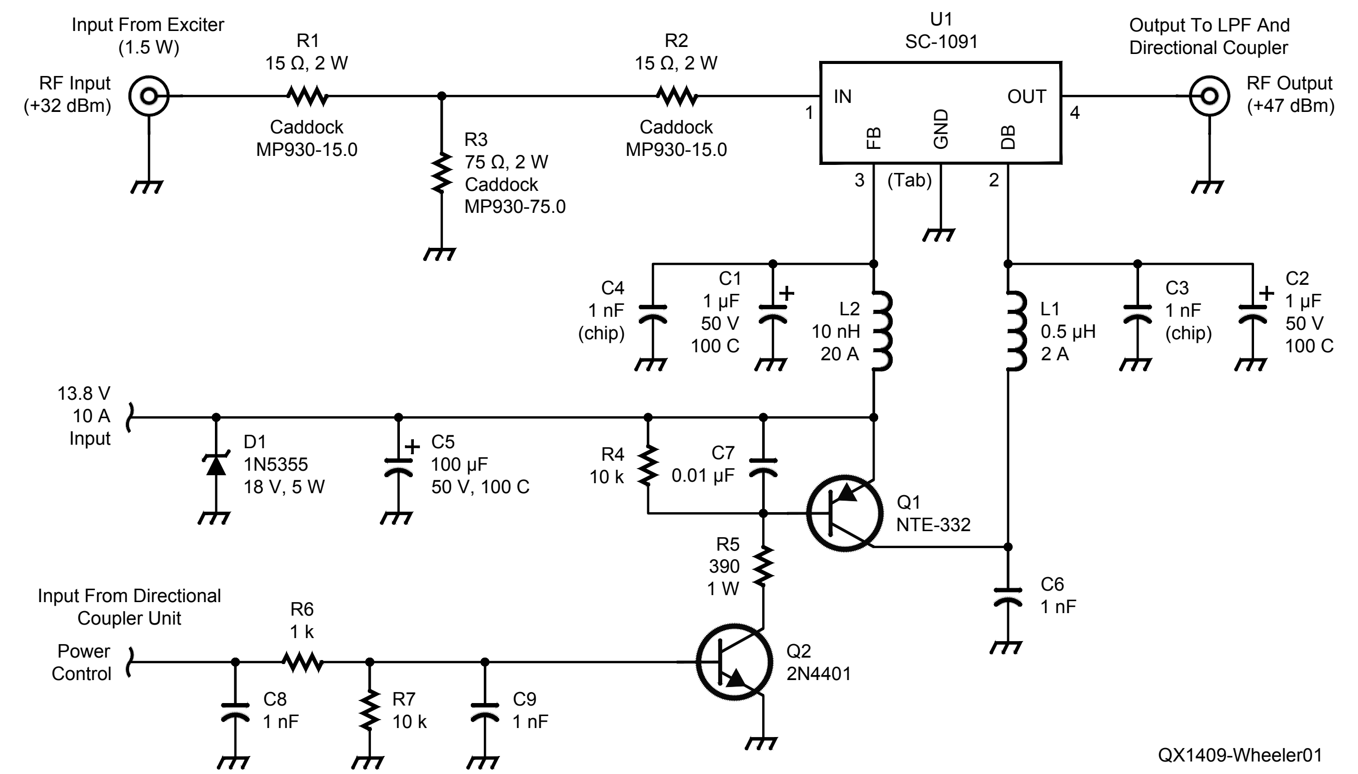


Figure 1 — Schematic for the 50 W MSR-2000 Replacement Power Amplifier Unit

separate stages (controlled gain stage, predriver, driver, and final) to amplify the 1 W output from the exciter to the 100 W final output level. These transistors are no longer available — and because of the age of these units, other components are just as likely to fail. Heat, time, and continuous duty cause passive components, particularly capacitors, in these units to fail in some very strange ways. This makes it difficult and time-intensive to repair these modules.

Fortunately, it's relatively straightforward to rebuild the amplifiers using modern thick-film ("brick") power integrated circuits. A single VHF power module such as the Toshiba SC-1091 can supply more than 50 W of continuous RF power with high reliability. These modules can easily be interfaced to the existing MSR-2000 control system, resulting in a "new" PA unit with high reliability and serviceability. Figure 1 is the schematic of a 50 W power amplifier replacement unit for the MSR-2000.

Circuit Analysis

The power amplifier is built around a single Toshiba SC-1091 integrated circuit. This module was chosen because it's rated at 60 W maximum output, and can be conservatively operated at the 50 W level. The IC has two class C amplifier stages — a driver and a final — each with its own collector power supply "B+" terminal (labeled DB for the driver stage and FB for the final stage). The driver stage is used as the controlled-gain stage by varying its B+ voltage.

The RF signal from the exciter first passes into a 5.4 dB attenuator T-pad built with noninductive power resistors, then into the IN terminal of the SC-1091. The 1.5 W maximum output of the exciter is reduced to about 430 mW maximum by this network in order to meet the input requirements of the SC-1091, which only requires about 300 to 400 mW of drive for full output.

The existing directional coupler and DC amplifier provides a Power Control signal for the power amplifier that becomes increasingly positive when more power output is required. This signal is phase-inverted by Q2 then passed to Q1, which then varies the voltage to the DB pin on U1 to control the amplifier power output. The DB pin should measure between 8 and 10 V DC when the amplifier is at full output. If excessive power is being produced, or excessive reflected power is present, the directional coupler unit will reduce the signal to Q2, thereby reducing the drive to U1 and decreasing the power output. The output of U1 passes directly to the existing low-pass filter assembly in the unit, and then through the directional coupler to the antenna.

When the transmitter is not keyed, the Power Control signal is at ground potential, which turns off Q2 and Q1, and also removes all bias from the DB pin on U1.

Diode D1 is incorporated for light surge protection, because the 13.8 V power supply in the MSR-2000 mainframe is not electronically regulated and may be subject to high voltage surges that can damage U1. Inductors L1 and L2, and capacitors C1, C2, C3, and C4 provide RF filtering for both the DB and FB power supply lines.

Construction

Figure 2 shows the construction of the unit. The circuit board was fabricated with old-school hand-layout of etchant-resist tape. Prior to fitting components onto the PA unit, the surface was faced flat using a mill, and mounting holes for the board and IC were threaded using a 6-32 tap.

To begin building this unit, remove all existing electronics from the assembly. Keep the directional coupler (left side of Figure 2) and input connection box (Figure 2, right) intact, because you'll be reusing these two parts.

It's important that the surface where the SC-1091 IC is mounted is kept very flat, preferably within 0.001 inch. Excessive ripple in the heat sink surface will cause the IC to flex slightly when its mounting bolts are tightened, resulting in a cracked ceramic substrate. Do not over-torque the mounting bolts, and use only enough heat sink compound to form a thin film on the bottom surface of the IC. If heat sink compound squishes out when you tighten the mounting bolts, you've used too much — and again, you run the risk of cracking the delicate ceramic substrate!

Resistors R1, R2, and R3 do not need heat sinks because they're 30 W units, and will be dissipating a maximum of ½ W each.

Runs carrying RF signals should be kept short, and liberal amounts of RF ground "pour" should be used to promote stable operation. Use chip capacitors everywhere possible, because they outperform leaded components at VHF and UHF. (It is quite possible to build the RF amplifier "ugly" style on a blank copper circuit board.)

A short run of 50 Ω Teflon coax connects the output of the board to the existing low-pass filter assembly through an RCA connector. Note that if you are going to use this amplifier design in a different application, you'll need to add a suitable low-pass filter to the output to meet FCC rules.

Transistor Q1 must have a heat sink; in the prototype, it's fastened to the main heat sink using a 6-32 nylon bolt and mica insulator.

You will need to transplant the input coaxial lead from the original amplifier board to the new RF amplifier board. This lead barely made the stretch, as you can see in Figure 2.

Splicing the board into the existing control system is quite simple. Just connect the Power Control lead from the directional coupler to the Power Control input on the amplifier unit. The remaining two leads from the directional coupler, B+ and Keyed A– should remain connected at their original locations on the power input and interface connectors. In the prototype you can see the directional coupler's orange B+ line is

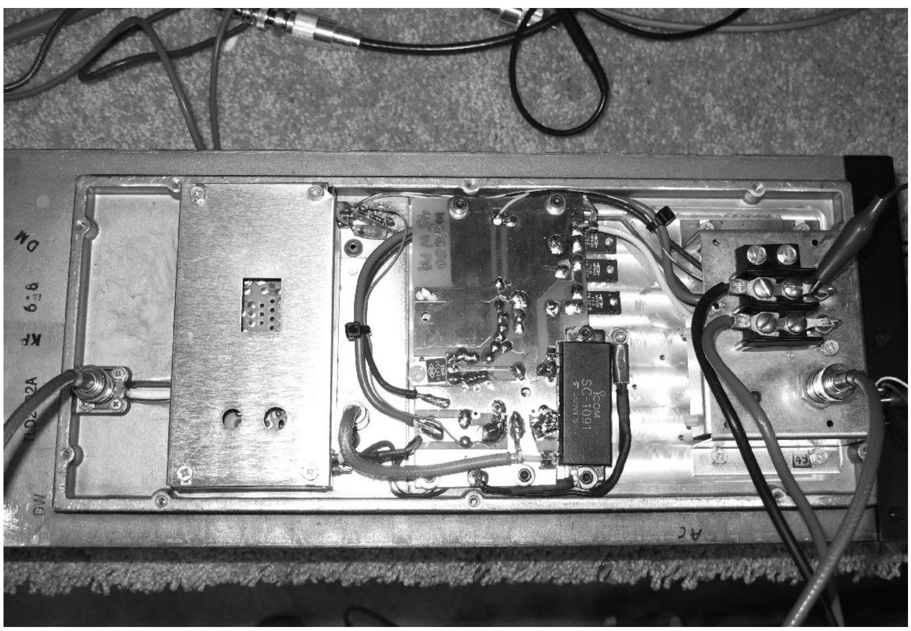


Figure 2 — Layout of the Power Amplifier

actually soldered to B+ on the circuit board.

Thermal Protection Circuitry

Our repeater is located at an outdoor site that experiences ambient temperatures in excess of 110°F on hot summer days. Inside the cabinet we've measured temperatures of 130°F and more under these conditions, and this is without the additional heat of the running system!

Thermal verification of the PA on the test bench using a Fluke thermal imaging system shows about a 70°F rise from ambient temperature (with no air flow) on the case of U1when the amplifier is operated at full power for five thermal time-constants (about 90 minutes), so with an ambient temperature of 130°F, the case of U1 could easily approach 200°F on a hot summer day and full power output. That's too close to the maximum limit of case temperature for U1 (212°F), so some additional protection is in order. Figure 3 shows the thermal protective circuitry.

A thermal switch with a cut-out temperature of 158°F is mounted very close to U1, and the leads to this switch pass back into the directional coupler unit. An additional 100 kΩ potentiometer is added to the directional coupler in series with the ground lead of R611 (just cut the ground trace of R611 and solder the new potentiometer, R610A, to the directional coupler circuit board to bridge the opened ground trace), and the thermal switch (normally closed) bypasses the 100 kΩ potentiometer. When temperatures rise above 158°F on the heat sink, this switch opens and power will be reduced according to the setting of R610A. Reducing the power to 5 or 10 W will provide plenty of safety margin in hot conditions, and allow the PA assembly to cool back down to a safe operating temperature.

Figure 4 is a thermal image of the amplifier under operating conditions. The instrument's cursor is positioned at the lower-left edge of U1. [This interesting color image doesn't really show the thermal gradient across the amplifier when printed as a gray scale image in *QEX*. We have placed the color image in the *QEX* Files section of the ARRL website.[1] — *Ed.*]

Testing

Checkout of the unit is very simple. Connect a suitable VHF wattmeter to the output connector on the assembly, a 12 A, 13.8 V supply, and a +30 dBm (1 W) signal to the input connector. Adjust R611 and R610 in the coupler unit for minimum resistance.

Leave the "A– key" signal on the interface connector unconnected. No RF output should be present, and the DB signal on U1 should be at zero volts DC.

Ground the A– key signal; RF output should be present, and you should be able to adjust R611 in the coupler unit for 45 to 50 W of output. You may need to nudge R610 slightly to accomplish this.

If the power output increases above 60 W (the IC can momentarily produce 70+ W of output, but it won't last long!), immediately unkey the unit and determine the problem.

Vary the input signal power over the range of 1 to 2 W. The RF output should remain steady at 45 to 50 W. This verifies that the power control circuitry is working correctly.

If you've added the thermal protection option, momentarily open of the leads to the thermal switch to simulate an overheat condition, then adjust R610A for 5 to 10 W of power output.

Conclusion

It's easy to rebuild the MSR-2000 VHF repeater power amplifier with modern components. This approach is a bit intensive

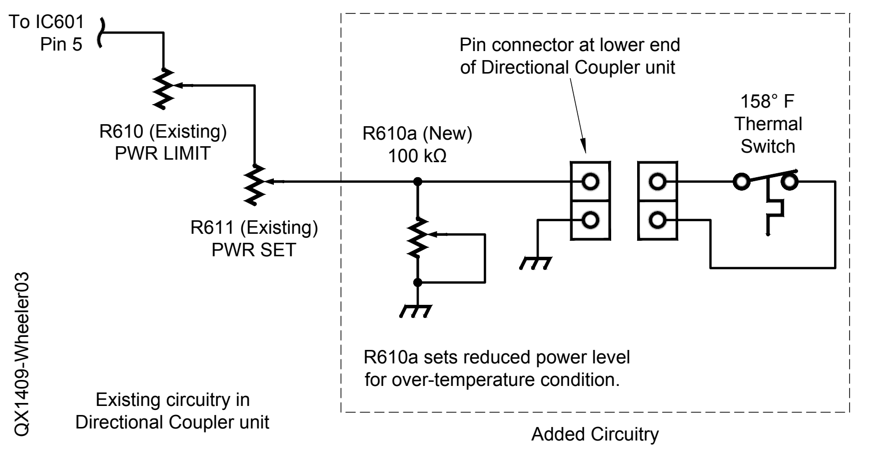


Figure 3 — Thermal Protective Circuitry

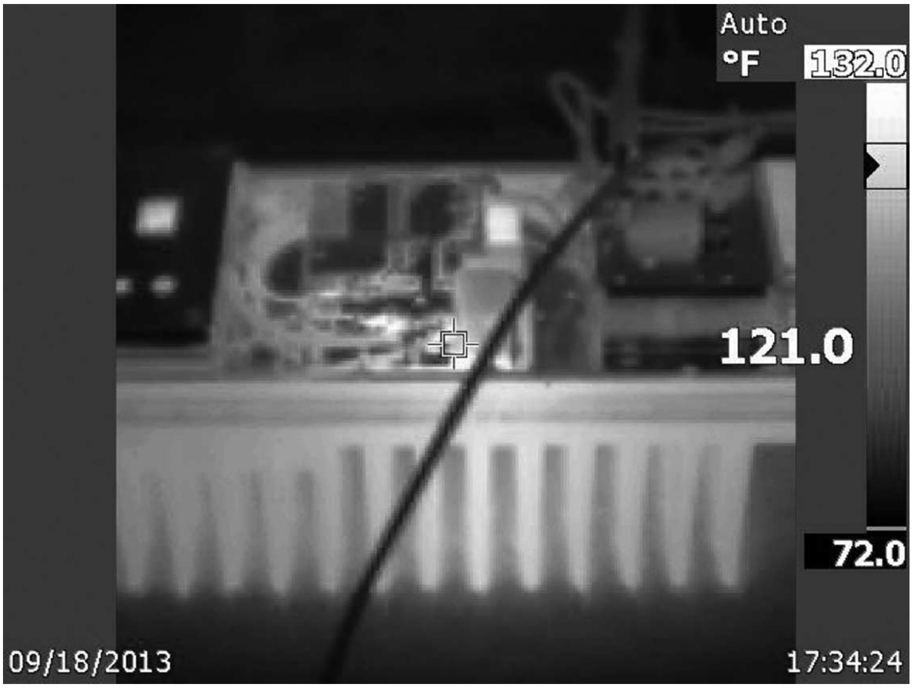


Figure 4 — Thermal Image of Operating Amplifier (P=50 W, Ta=74 F, Ton=90 minutes)

[1]The full color image for Figure 4 is available for download from the ARRL QEX Files website. Go to **www.arrl.org/qexfiles** and look for the file **9x14_Wheeler.zip**.

mechanically because of the required hand-fitting and machining work, but the end result is a "like new" PA that should provide years of high-reliability operation for your repeater system.

Acknowledgement

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*Tom Wheeler has been licensed as NØGSG since 1985. An ARRL Member, Tom is Dean of Instruction at the Metropolitan Community College Business and Technology Campus in Kansas City, Missouri (**www.mcckc.edu/btc**). He is the author of the Prentice Hall textbook* Electronic Communications for Technicians *(now in its second edition), and is active in the Johnson County (Kansas) Radio Amateurs Club (**www.w0erh.org**).*

Tom's technical interests include analog and digital signal processing, RF design, and computer languages. He holds Associates and Bachelor's degrees in Electronic Engineering Technology, a Master's degree in Technology, with emphasis in digital controls and signal processing. Tom also has a Doctorate in Education.

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Colin Horrabin, G3SBI

17 Denbury Ave, Stockton Heath, Warrington, WA4 2BL, Unite Kingdom; **colin.horrabin7@ntlworld.com**

The Development of the Low Phase Noise Double Tank Oscillator

The author discusses local oscillator phase noise issues for receiver performance, and describes his efforts with a double tank oscillator circuit.

Prior to the design of low cost frequency synthesizers for use as the local oscillator in a super heterodyne HF/LF receiver, most tunable oscillator designs were developed for their good frequency stability. A particular favorite of mine for home brew equipment in the early 1960s was the Vacker Oscillator, in which the transistor in the oscillating circuit had a 0.001 µF capacitor to ground from the base and from the collector, reducing the effect on frequency stability of the transistor parametric capacitances.

In Amateur Radio circles at the time you would not hear much talk of how local oscillator sideband noise (phase noise) could limit a super heterodyne receiver dynamic range because of an effect called reciprocal mixing. At the time, some high stability oscillator designs did have low sideband noise and some didn't. Few, if any, people had the equipment to measure it. If you were operating in a contest with plenty of big signals on the band, however, you could quantify practical differences in oscillator performance by the radio's big signal handling.

In full frequency synthesis, a voltage controlled oscillator (VCO) would be the

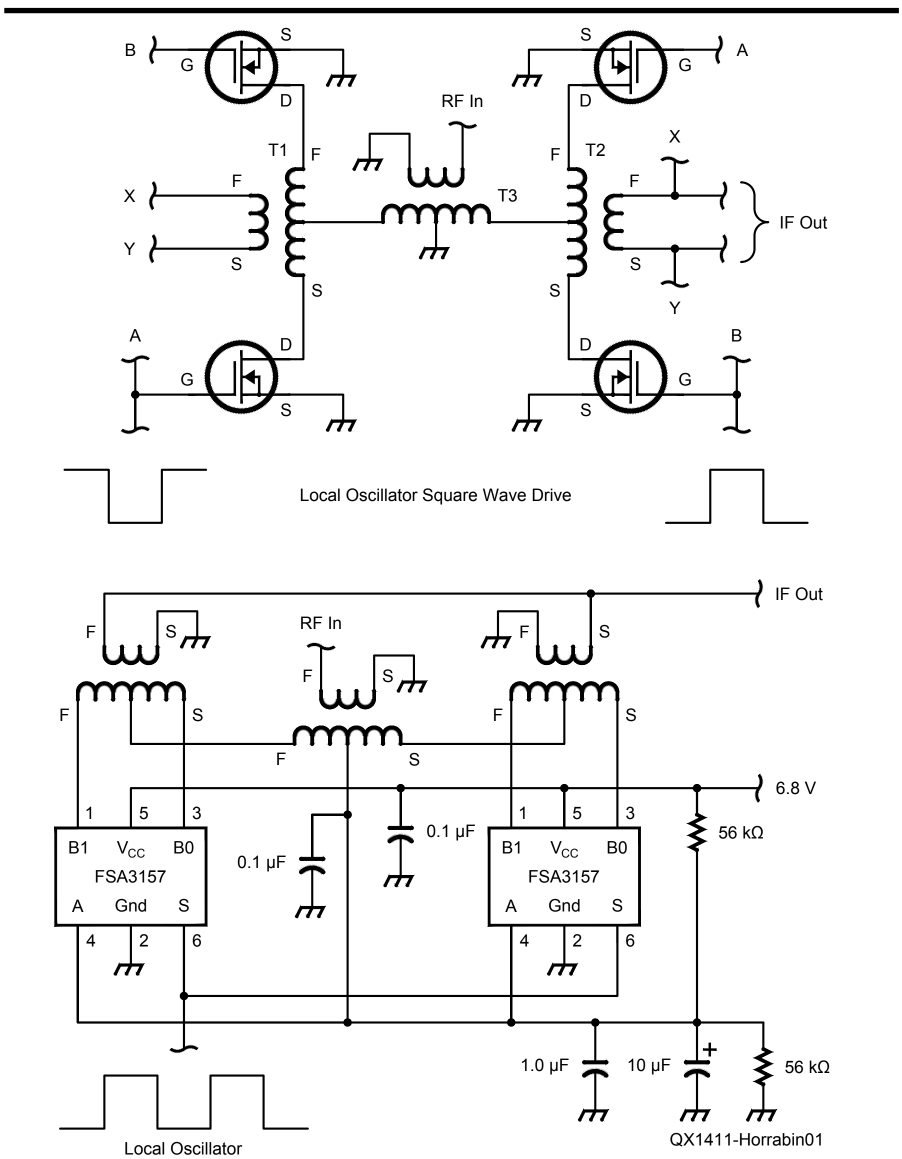

Figure 1 — Part A is the H Mode mixer as originally sketched on the back of an envelope. This required complementary square wave drive from the local oscillator. It was called the H Mode mixer because of the way the transformers in the sketch formed the letter H. Part B shows the H Mode mixer as it was used in the HF7070 receiver, configured to use the Fairchild FSA3157 fast bus switch. The FSA3157 is an SPDT switch with 0.5 ns break before make action, so the local oscillator drive logic does not have to generate a square wave plus its complement.

local oscillator and this would be phase locked with digital logic to a reference crystal oscillator so that the frequency stability was that of the quartz crystal and not of the VCO. In practice other features of the local oscillator performance then became an issue such as reference frequency modulation of the VCO and low level birdies for various reasons in multi loop frequency synthesizers. The VCO itself would be part of a phase locked loop so outside the loop bandwidth (1kHz or so) the VCO sideband noise was that of the VCO itself which after a certain point should fall at 20 dB/decade with offset frequency from the VCO carrier frequency.

In 1993 Pat Hawker, MBE, G3VA (SK), introduced me to an excellent article about high dynamic range front ends for HF receivers that had been published in the February 1993 issue of *QST*. The author was Jacob Makhinson, N6NWP.This article really interested me and resulted in my development of the H-Mode mixer. See Figure 1. This mixer used the Siliconix SD5000 DMOS FET array, which with square wave drive gave an input IP3 above 50 dBm on the HF bands. This development was reported in the October 1993 issue of *RadCom*, in the Technical Topics column. It turned out that if sine wave local oscillator drive was used the mixer IP3 was only 35 dBm and that you needed square wave drive to get 50 dBm. Jacob Makhinson, N6NWP, had solved this problem rather neatly by using a 74HC74 bistable type D flip-flop on a 9 V supply as a divide by two squarer. The real problem was that using the SD5000 array in the H-Mode mixer you really needed a spectrum analyzer to set it up, and as a result I don't think it was used in any Amateur Radio projects at that time.

The situation changed in 1998 when Gian Moda, I7SWX, spotted an announcement that Fairchild had introduced the FST3125 fast bus switch, and when used as an H-Mode mixer it could be driven by ordinary logic. Bill Carver, W7AAZ, managed to get some samples within a few days of this news and he constructed an H–Mode mixer based on this chip with excellent results. He achieved an IP3 of 45 dBm, and this development was reported in the Technical Topics column of *RadComs*. As a result, Bill's circuit has subsequently been used in a number of Amateur Radio projects to give the receiver third order intercept points above 40 dBm.

The original October 1993 write up on the H-Mode mixer was followed in the January 1994 *RadCom* Technical Topics column on terminating the mixer with quadrature hybrid connected 4 pole roofing filters of 2.5 kHz bandwidth. At 9 MHz, the total losses involved were only 1.1 dB, so that a sensitive high dynamic range receiver could be made without a preamplifier before the H-Mode mixer.

The January 1994 Technical Topics column also included a letter from Peter Chadwick, G3RZP, a former RSGB President and a principal engineer at Plessey Semiconductor. In his letter Peter said that in practice, the receiver signal path dynamic range resulting from the use of an H-Mode mixer would be seriously limited by reciprocal mixing from local oscillator sideband noise (phase noise). He pointed out that the designer was in the classic situation of improving one thing and then having to improve something else.

This was something I was already aware of and had decided to look at basic oscillator circuits to see if one could be found with phase noise that fell away at a rate greater than 20 dB/decade from the carrier. The most likely way to achieve this was to look at oscillators that used more than one resonator in the oscillating circuit. The reason for this approach is that in the design of a band-pass filter, as you add more sections the skirts get steeper with offset frequency. Therefore, if two tuned circuits could be part of an oscillator design you would expect phase noise to fall at a rate greater than 20 dB/decade from the carrier.

The G3PDM Receiver

In the early 1970s quite a few radio amateurs in the UK built the G3PDM receiver. It appeared in a series of articles in *RadCom* in 1971 under the title of "Hybridise and Plagurise" and it seemed to be good at big signal handling. The local oscillator circuit used partial frequency synthesis, where the local oscillator VCO used two triode tubes and was mixed against a crystal oscillator and then phase locked to a Vacker VFO running in the 5 MHz region. This design had a very narrow PLL loop bandwidth, so there had to be something special about the basic phase noise profile of the VCO itself to account for

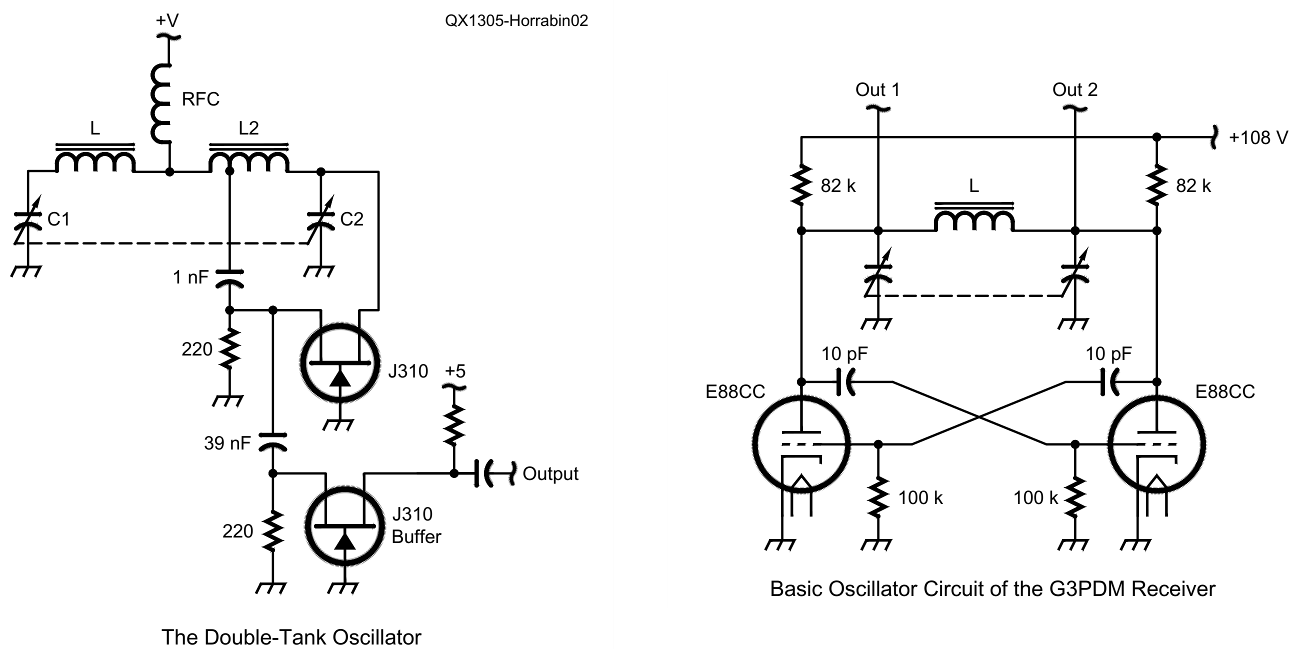

Figure 2 — Part A shows the HF7070 double tank oscillator. Part B shows the tube-based oscillator used in the G3PDM receiver, designed by Peter Martin in 1970.

the radio's excellent big signal performance

The basic G3PDM circuit is shown in Figure 2. Apart from the fact that there are two tuning capacitors, there is nothing obvious to show why it should have a good phase noise profile, other than that two ganged variable capacitors were needed to tune the circuit. In 1994 it seemed impossible to build that circuit using JFETs because of the lower impedances involved in solid state electronics compared to tubes. I looked at another circuit that should behave in a similar manner however, because two ganged capacitors were needed to tune it and this became known as the double tank oscillator (Figure 2).

Superficially, it is easy to understand how the double tank oscillator could work because two resonators are involved. It's not so easy from a good phase noise point of view to understand how G3PDM's VCO could work. To understand this circuit it is best to get rid of one of the triodes to make the circuit single ended (Figure 3) and split the coil into two coils. On the left you have a series tuned circuit in series with L2, which is part of a parallel tuned circuit driven by the anode of the triode. At resonance, the series tuned circuit only presents its resistance in series with the inductor, L2. This reduces the *Q* of the parallel tuned circuit by a factor of two. As you move away from the carrier frequency however, two tuned circuits are active. Therefore this should increase the rate of fall off caused by oscillator phase noise beyond 20 dB/decade.

The real problem in 1994 was whether any circuit simulation package of the day could be used in some way to indicate the relative phase noise performance of any oscillator circuit. If it could, and a basic oscillator circuit was found that was significantly better in some way compared to a single resonator oscillator, it would then have to be built and tested. This was not an easy job at the time, with the test equipment that was then available in the RF Lab at Daresbury Laboratory.

Micro-Cap 3 Circuit Simulation Software

Daresbury Laboratory had a copy of *Micro-Cap 3*, which operated under DOS on a PC. I had become an expert at using this software package. As part of the development of a high dynamic range front end for a down conversion receiver, I had used this package to simulate a 4 pole ladder filter of 2.5 kHz bandwidth at 9 MHz.

For this particular application an equivalent circuit of the 9 MHz crystal was obtained by using a Hewlett Packard 4195A spectrum and vector network analyzer. The crystal parameters were then used in the simulation of the 9 MHz filter. The simulation showed a 0.9 dB insertion loss for the 4 pole 2.5 kHz bandwidth filter, which seemed remarkably low at the time. The practical design was identical in every way except that the insertion loss was even lower.

There was a reason for this. When measuring the impedance of a component using the HP 4195A and you selected the "more" function it gave you a choice of calculating the parametric values with frequency of capacitors, inductors and a quartz crystal. When you used the "more" function there seemed to be a software bug, because for an inductor, its series resistance was always a factor of two higher than it should be. This same problem also occurred with crystals. If you wanted to know the series R value of a coil at a particular frequency, the best way was to turn it into a series tuned resonant circuit to get the R value. This in no way spoiled the fact that the HP4195A was a superb development tool for the RF engineer, and I was later able to use it as part of my oscillator phase noise measurement set up.

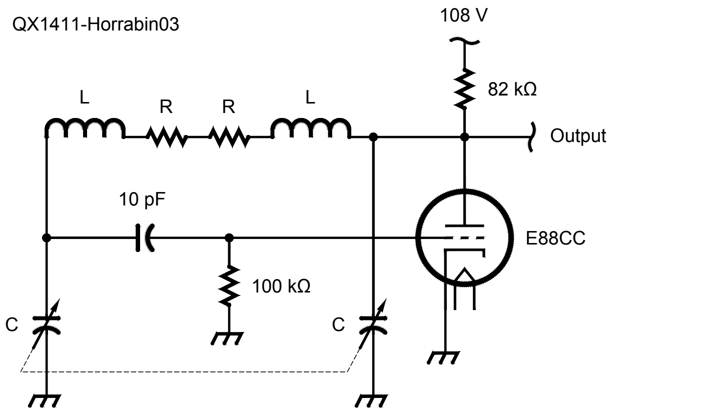

Figure 3 — This is the equivalent circuit for a single ended G3PDM oscillator, to help study the implications of the single and double tank circuit from a phase noise point of view. The single coil shown in Figure 2 is split into two coils. This creates a series tuned circuit in series with the inductor of a parallel tuned circuit. At resonance, the series inductor just presents its series resistance, reducing the parallel tuned circuit *Q*. As you move away from the resonant frequency, however, the two resonant circuits are active, which should increase the fall off of the phase noise as you move away from the carrier, as compared to a conventional oscillator.

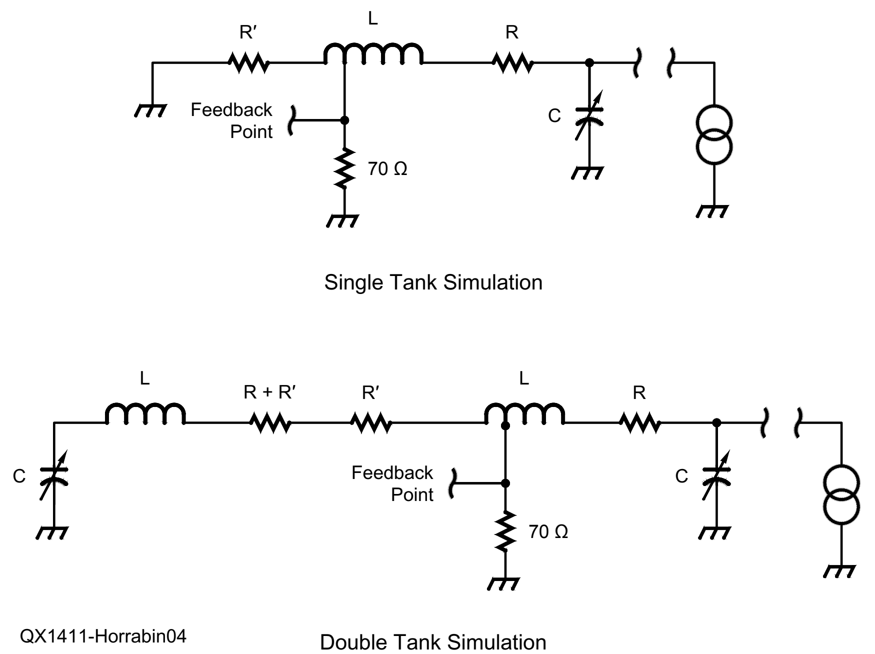

Figure 4 — These two circuits were used in the *Micro-Cap 3* simulation. The parallel resonant frequency was swept using an AC current source. The rate of change of phase with offset frequency was plotted across the 70 Ω resistor, representing the input resistance of a J310 transistor operating in grounded gate configuration.

Quantifying Relative Oscillator Phase Noise Performance Using *Micro-Cap 3*

Initially I decided to use *Micro-Cap 3* to analyze the two basic circuits shown in Figure 4. Basically, *Micro-Cap* was used to frequency sweep across the parallel resonance of these circuits and look at the rate of change of phase at the oscillator feedback point.

The L, C, and R values used in the simulation gave a coil *Q* of about 70, and were resonant in the 50 to 60 MHz region. The thinking behind this was that for a down conversion receiver you would divide this frequency by ten to get a further improvement in phase noise, so your VFO would be running from 5 to 6 MHz.

In this simulation the double tank circuit had a phase rate of change that was a factor of ten greater at the oscillator feedback point than the equivalent single tank oscillator. To see if this simulation was a valid method to indicate the relative phase noise profile of a circuit, a double tank oscillator had to be built and its phase noise profile measured. At the time, the same techniques were not applied to the G3PDM oscillator circuit. That was a major failing on my part. It was just that I didn't see it as relevant for a solid state oscillator design because of the lower impedances involved.

The Construction of an Oscillator Phase Noise Measurement System

In 1994 I knew very little about the techniques to measure oscillator phase noise. John Thorpe, who then worked for Lowe Electronics in Matlock Derbyshire was well known in the UK as the designer of the "Lowe" receivers. These were budget priced up-conversion receivers for the discerning short wave listener. It is to John's credit that they were beautifully engineered

I contacted John and asked him how he was measuring phase noise. It turned out that he had no equipment to measure it directly at this time but that he could calculate it from his receiver reciprocal mixing measurements. In any receiver the measurement of reciprocal mixing is what really matters and not the direct measurement of your local oscillator phase noise. This is because even if your oscillator had excellent phase noise, this could be destroyed by the circuitry before the mixer. If you are developing an oscillator, however, you need to be able to measure phase noise directly to speed up the development process so I put together a system to do this. See Figure 5.

In 1994 a good spectrum analyzer used to look at oscillator sidebands would have a 100 dB dynamic range. This would need to be extended by at least 30 dB to show low levels of oscillator phase noise relative to the carrier level. The accepted technique was the quadrature lock method to notch out the carrier by up to 40 dB followed by a 40 dB gain low noise amplifier to effectively increase the spectrum analyzer dynamic range to display low level oscillator sidebands.

The signal whose phase noise we wanted to measure went to the RF input of a mixer. The local oscillator input to this mixer came from a reference oscillator that had to have a phase noise performance at least as good as the oscillator to be measured. Feedback of the DC voltage output from the mixer after a 40 dB gain low noise amplifier would go to the DC FM input of your reference oscillator to hold it in phase quadrature with the oscillator to be measured. The bandwidth of the DC feedback was below 1 Hz. This circuit reduced the amplitude of the carrier by about 40 dB, but left the amplitude of the sidebands above a few Hertz unchanged. The sidebands were then amplified by 40 dB in a low noise DC coupled amplifier with a bandwidth of 1 MHz before being displayed on a low frequency spectrum analyzer.

The bandwidth of the closed loop was defined by the 560 Ω and 56 kΩ resistor network shown in Figure 5. The 560 Ω resistor was necessary to get the feedback system in lock. This was because if the lock was leading by 90º rather than lagging, you could have positive feedback. This resistor network helped to establish the lock for negative feedback, to obtain a stable situation.

The spectrum analyzer side of the HP4195A was used to display phase noise against carrier offset frequency. Some of the features on the HP4195A really came into their own with this measurement technique. For a start, you could select a logarithmic frequency scale. Also, you could select the display of amplitude as a dB/Hz scale. This was really useful because as you changed the resolution bandwidth on the spectrum analyzer, the position of the phase noise plot did not change. So you could do a fast sweep with 100 Hz bandwidth to get a general plot and then you could follow it with a much slower plot using 3 Hz bandwidth to show more detail particularly very close to the carrier.

It was possible to confirm that the double tank oscillator phase noise did fall at 30 dB/decade. The oscillator circuit details were shown in *RadCom* Technical Topics in 1994 but because I was using home made test gear I did not think I could make specific claims as to the oscillator phase noise profile.

It became necessary to wait until Peter Hart, G3SJX, did an independent review of the AR7030 receiver in 1996, which used John Thorpe's version of the double tank oscillator, to confirm the performance of the double tank VCO in that receiver. At first glance the phase noise slope of 30 dB/decade from this circuit overturns Leeson's Equation. Maybe it doesn't if, for some reason, it increases the flicker noise corner frequency of the J310 FET. That is why it would be interesting if a mathematician investigated the double tank circuit in some detail.

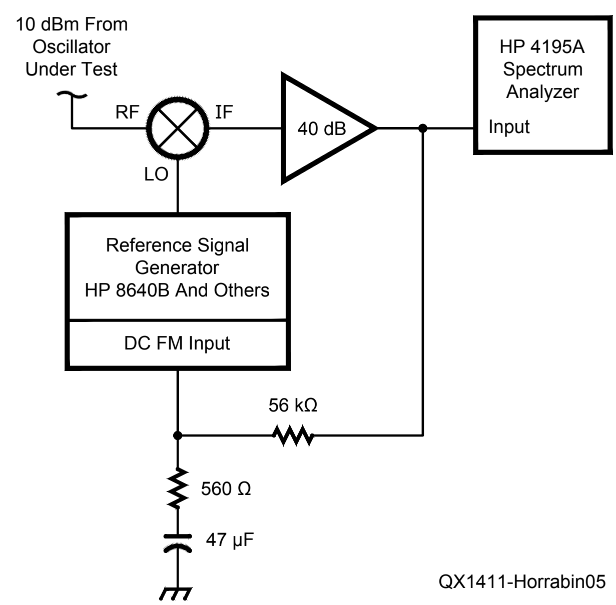

Figure 5 — This diagram illustrates the phase noise measurement system.

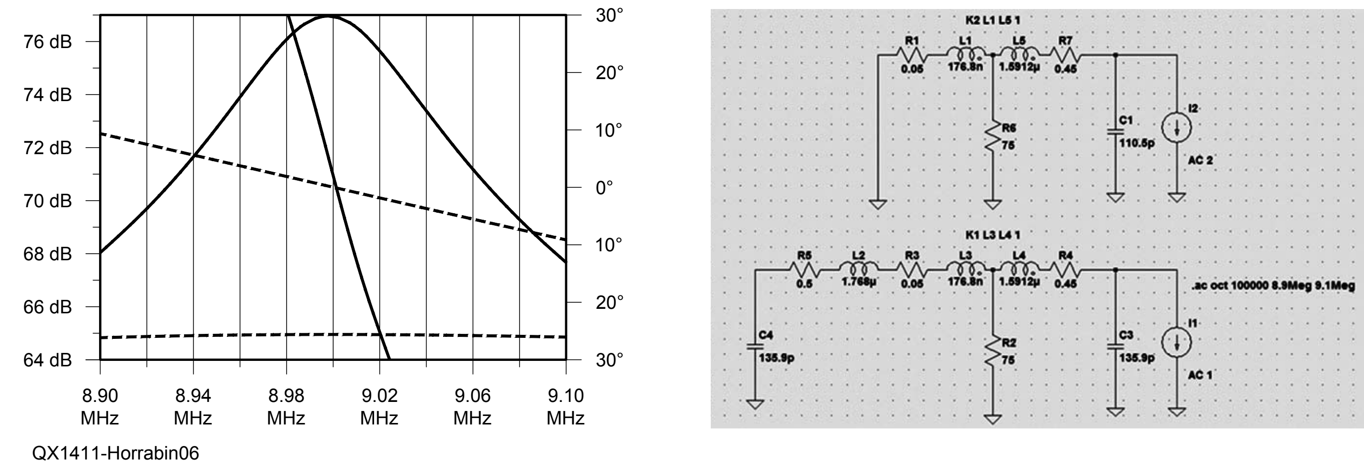

Figure 6 — The curve with peak response just above 9 MHz represents the voltage across the inductor closest to the AC source for the G3SBI double tank circuit. The sloping line that goes from about 72.5 dB to about 69 dB represents the voltage across the circuit from the tap between the inductors to the AC source. Part B shows representations of the *Spice* circuit model.

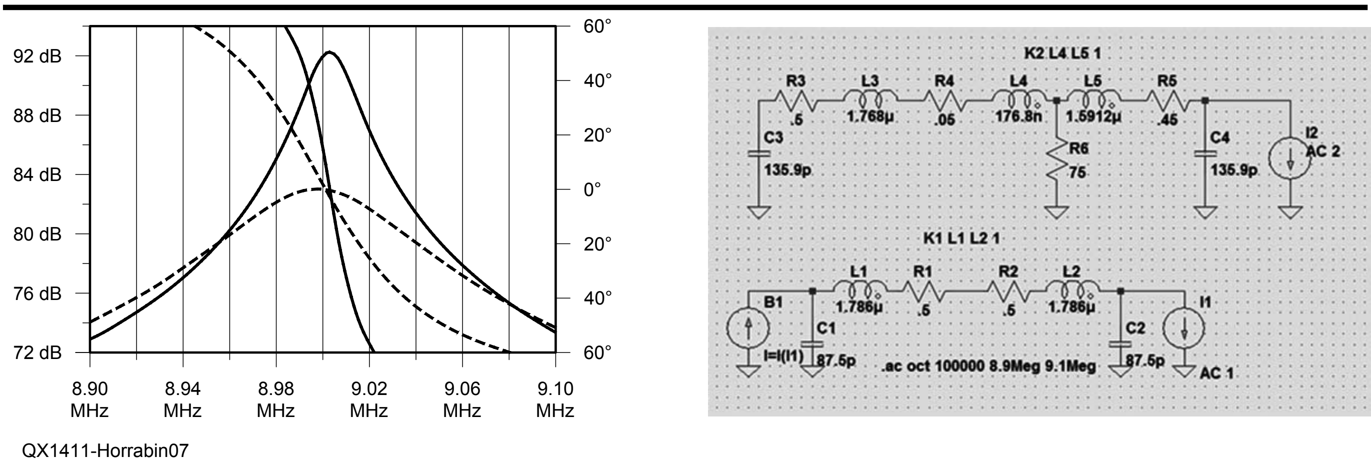

Figure 7 — The curve that peaks at about 93 dB represents the voltage across the left-most inductor in the circuit simulation. The curve that peaks at about 83.5 dB represents the voltage across the right-most inductor. Part B shows representations of the *Spice* circuit model.

From Wikipedia (**en.wikipedia.org/wiki/Leeson's_equation**), we learn that Leeson's Equation is an empirical expression that describes an oscillator's phase noise spectrum. Leeson's expression for single-sideband (SSB) phase noise in dBc/Hz (decibels relative to output level per Hertz) is:[3, 4]

$$L\left(f_m\right) = 10\log\left[\frac{1}{2}\left(\left(\frac{f_0}{2Q_l f_m}\right)^2 + 1\right)\left(\frac{f_c}{f_m} + 1\right)\left(\frac{FkT}{P_s}\right)\right]$$

where:
f_0 is the output frequency
Q_l is the loaded Q
f_m is the offset from the output frequency (Hz)
f_c is the $1/f$ corner frequency
F is the noise factor of the amplifier
k is Boltzmann's constant
T is absolute temperature in Kelvins
P_s is the oscillator output power.

Circuit Simulation by PA3AKE

Even though the double tank oscillator was designed nearly twenty years ago, its virtues seem to be unknown to mainstream radio designers. That is one reason I have written this article about its development.

The plots that were made of phase slope at the oscillator feedback point using *Micro-Cap 3* were long since lost, so I asked my friend Martein Bakker, PA3AKE, if he would repeat these plots for this article using *Spice*.

Martein obtained some very interesting results, which will be described in the following paragraphs. He also did a simulation for the oscillator in the G3PDM receiver. That is something that I wish I had done 20 years ago

On his website, Martein describes the three different methods he used to measure sideband noise on the AD9910 DDS chip.[1] This includes the method of using of a crystal notch filter to take out the carrier, which was described by Wes Hayward, W7ZOI, in a Jul/Aug 2008 *QEX* article.[2] Although this method will not allow you to measure phase noise very close to the carrier like the quadrature lock method, it has two distinct advantages.

First, you don't need a reference oscillator. Second, unlike the quadrature lock method, it will also respond to AM noise on the carrier.

In their earlier DDS chips Analog Devices provided a pin to bypass low frequency noise from an op amp on the chip, which set the reference current to the high speed DAC. They didn't do this with the 1 GHz parts because the Agilent Signal Source Analyzer they were then using to measure phase noise used the quadrature lock method. This suppressed the effect of

AM noise by about 40 dB, so they didn't see the noise. Of course they could have bought an AM noise measuring option for the magic Agilent box, but they didn't. It is a mistake they won't be making again. It does show that you need to know the limitations of your test equipment, however, especially for low noise measurements that are approaching the thermal noise floor.

The 1 GHz DDS chips designed by Analog Devices are a major technical achievement (even more so is the 3.5 GHz AD9914) but the 1 GHz chips have an AM noise issue. Analog Devices was not aware of this problem until Martein became involved. There is a write up of his investigation of the AD9910 as a local oscillator for his holy grail down conversion receiver on his website (see Note 1). Ultimately, using a low phase noise 1 GHz clock and an ultra low noise voltage regulator for the AD9910, Martein was able to get within 1 dB of the AD9910 residual phase noise measurements shown in the AD9910 data sheet.

Simulation of the Double Tank Oscillator Phase Slope by PA3AKE

When I gave Martein the equivalent circuits of the single and double tank oscillator to test, I forgot the 70 Ω resistor to ground to simulate the input resistance of the J310 FET in a grounded gate configuration. When Martein did the plot there was no difference between the single and double tank circuits. With the 70 Ω resistor added to the circuit, he obtained the same results I had twenty years ago. That JFET input resistance is an important part of the model!

The two coils used in the double tank circuit are not mutually coupled, so it would appear that coupling to the dummy tank occurs via the input impedance of the amplifying device. This was something I had not realized. The single and double tank phase slope graphs are shown superimposed in Figure 6. Note that the double tank slope is a factor of ten greater than the single tank oscillator.

Simulation of the G3PDM Oscillator's Phase slope by PA3AKE

The phase slope graph of the G3PDM oscillator is shown in Figure 7 superimposed on that of the double tank oscillator. The phase slope is a factor of 2.5 times better than the double tank oscillator. So the G3PDM oscillator should give a phase noise profile of 34 dB/decade, based on the double tank measured performance of 30 dB/decade.

The Local Oscillator VCO in the Drake TR7 Transceiver

Martein owns a Drake TR7 transceiver, and he had a surprise when he first saw the detailed circuit of the G3PDM VCO. Although there were minor circuit differences, it seemed that Drake had made a JFET version of the G3PDM VCO for use in the TR7. According to the Sherwood Engineering list (**www.sherweng.com/table.html**), its phase noise at 10 kHz of –116 dBc/Hz was nothing special. Martein has now measured the phase noise profile of his TR7 on the 7 MHz band. These results are shown in Table 1. The phase noise profile of the TR7 is particularly interesting, and this shows the shortcomings of quoting phase noise at one particular offset frequency, like in the Sherwood list.

Table 1 shows the phase noise profile in dBc/Hz on the 7 MHz band of the local oscillator for the Drake TR7 and the HF7070 Receiver. The TR7 has a phase noise plateau of –117 dBc/Hz up to 20 kHz, which suggests that the PLL bandwidth is around 20 kHz. Although it does not show in Table 1, the phase noise profile of the HF7070 up to its PLL bandwidth of 1 kHz was –116 dBc/Hz. Both the HF7070 and the TR7 use a similar phase detector chip, so this chip appears to impose a phase noise limit within the PLL bandwidth.

Outside of the PLL bandwidth, the local oscillator in the TR7 falls at 12 dB/octave. This is the magic 40 dB/decade for a G3PDM type oscillator. Reducing the PLL bandwidth to 1 kHz on the TR7 could have made the receiver on the TR7 a state of the art performer. Running the oscillator open loop (unlike the HF7070) the TR7 VCO has more noise very close to the carrier.

In this situation, if the PLL bandwidth

Table 1
Phase Noise Profile

Offset kHz	*5*	*10*	*20*	*30*	*40*	*50*	*100*	*200*
TR7	–117	–117	–117	–126	–131	–136	–146	–153
HF7070	–126	–138	–147	–150		–154	–159	–162

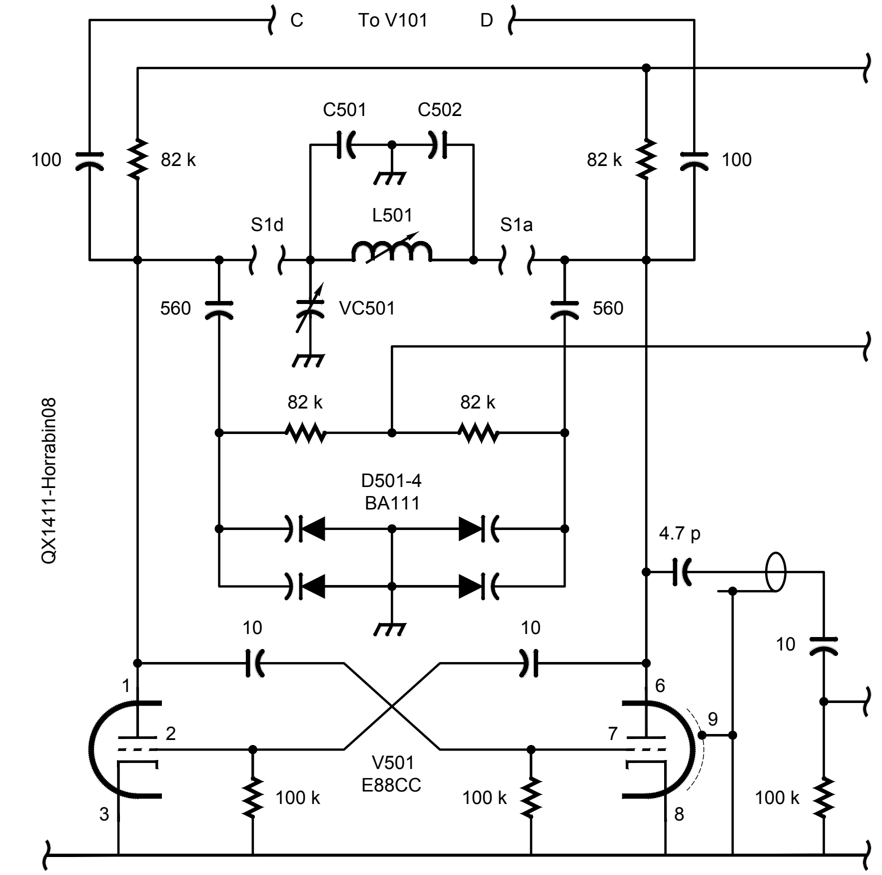

Figure 8 — This schematic diagram shows the G3PDM receiver VCO.

in the TR7 had been reduced to 1 kHz, the phase noise would have been unacceptably poor at an offset frequency of 1 kHz. That could explain why the PLL bandwidth was set at around 20 kHz. This would give a more reasonable result at 1 kHz offset, because of the extra loop gain at 1 kHz for a PLL bandwidth of 20 kHz. By doing this, the phase noise at 1 kHz would be closer to that of the reference frequency phase noise in the PLL. The reference input would have a better phase noise at 1 kHz than that of the VCO open loop.

The reason the close in noise for the TR7 is higher than the double tank VCO is probably because of noise from the 78L05A regulator chip, U501, that supplies the VCO. See Figure 9A. The regulator is followed by an RC network using a 33 Ω resistor and a 47 μF capacitor to reduce the noise from the regulator. This will have little effect on the close-in noise, however, because the 3 dB point is around 100 Hz. So if this chip was responsible for the higher close in noise, and was replaced by a much lower noise voltage regulator, the TR7 could have become the best in the business. Since the TR7 was one of the first up-conversion radios designed in the 1970s, it had the potential to become a top radio in terms of phase noise, and would still be very good by today's standards.

Referring to Figure 9B, there should also have been a capacitor across the 3.9 kΩ resistor (R528) that is used to control the PLL loop dynamics in amplifier "A." This capacitor would normally have been about a fifth of the value of C541, which is a 0.01 μF capacitor. So a capacitor of about 0.002 μF across R528 would have been in order. By doing this, it further rolls off the gain of op amp "A" above the PLL bandwidth, and should not cause loop instability. If it did, the value of the capacitor across R528 would have to be reduced in value.

Observations

Some people may think that the matching of varicaps used in a double tank VCO would be critical. This is not the case because the circuit itself compensates for small differences between the two varicaps. Around 5000 AR7030 receivers were made and about 800 CDG2000 transceivers have been constructed and this has never been an issue.

In principle, the G3PDM type oscillator VCO, using JFETs in the Drake TR7 could have been the best in the world even by today's standards with the phase noise falling at 40 dB/decade. An AD9910 DDS chip clocked by a low noise 1 GHz surface acoustic wave (SAW) oscillator, however, would give even better phase noise close in.

Bill Squires, W2PUL, designed the SS-1R receiver in the 1960s. This firmly established some new ideas for use in high dynamic range receivers. The G3PDM receiver designed by Peter Martin capitalized on the principles established by Bill Squires. Both of these receivers were designed before modern testing methods were introduced. As a result, today there is increased interest in these two receivers from those of us who are interested in high dynamic range receivers. Unfortunately, I don't think many of these receivers have survived because most people for some time now prefer to use modern commercial products — even those of us who are capable of designing our own gear.

A Google search on the SS-1R found an American amateur who had reconditioned an SS-1R to the manufacturer's specification, to test the radio. Likewise on this side of the pond an advertisement in *RadCom* for anyone who had a G3PDM receiver produced a result. Someone had bought an incomplete one on eBay and intends to complete it. Maybe this will find its way to Peter Hart, G3SJX, at some point to test. Of course if the G3PDM receiver had been subject to modern testing methods in 1971, its good reciprocal mixing measurements would have generated a lot of interest from main stream radio designers.

I would have thought that designers of commercial Amateur Radio equipment would read our Amateur Radio magazines, and would have found both the H-Mode mixer and the double tank oscillator, and made use of these concepts. Apparently not, because it would appear that the new Kenwood TS990s is the first commercial radio designed in Japan to use an H-Mode mixer. The details of the H-Mode mixer were placed in the public domain 20 years ago.

Acknowledgements

Both the H-Mode mixer and the double tank oscillator were developed in 1993 and 1994 when I worked in the RF group at Daresbury Laboratory. This experimental work would not have been possible without the agreement of the head of RF engineering, Mike Dykes. Both he and I are electrical engineering graduates of Salford University, in Manchester. When I first studied electrical engineering there it was still known as The Royal College of Advanced Technology, and was a four year "sandwich" course. It had the reputation of producing the best electrical and electronic engineers in the UK. To take this course as well as having the necessary academic entry qualifications, you had to be sponsored by a firm where you would do the necessary industrial training. In my case the British Aircraft Corporation at Warton Aerodrome (a former B17 base in WW2) where I was an apprentice, served as my

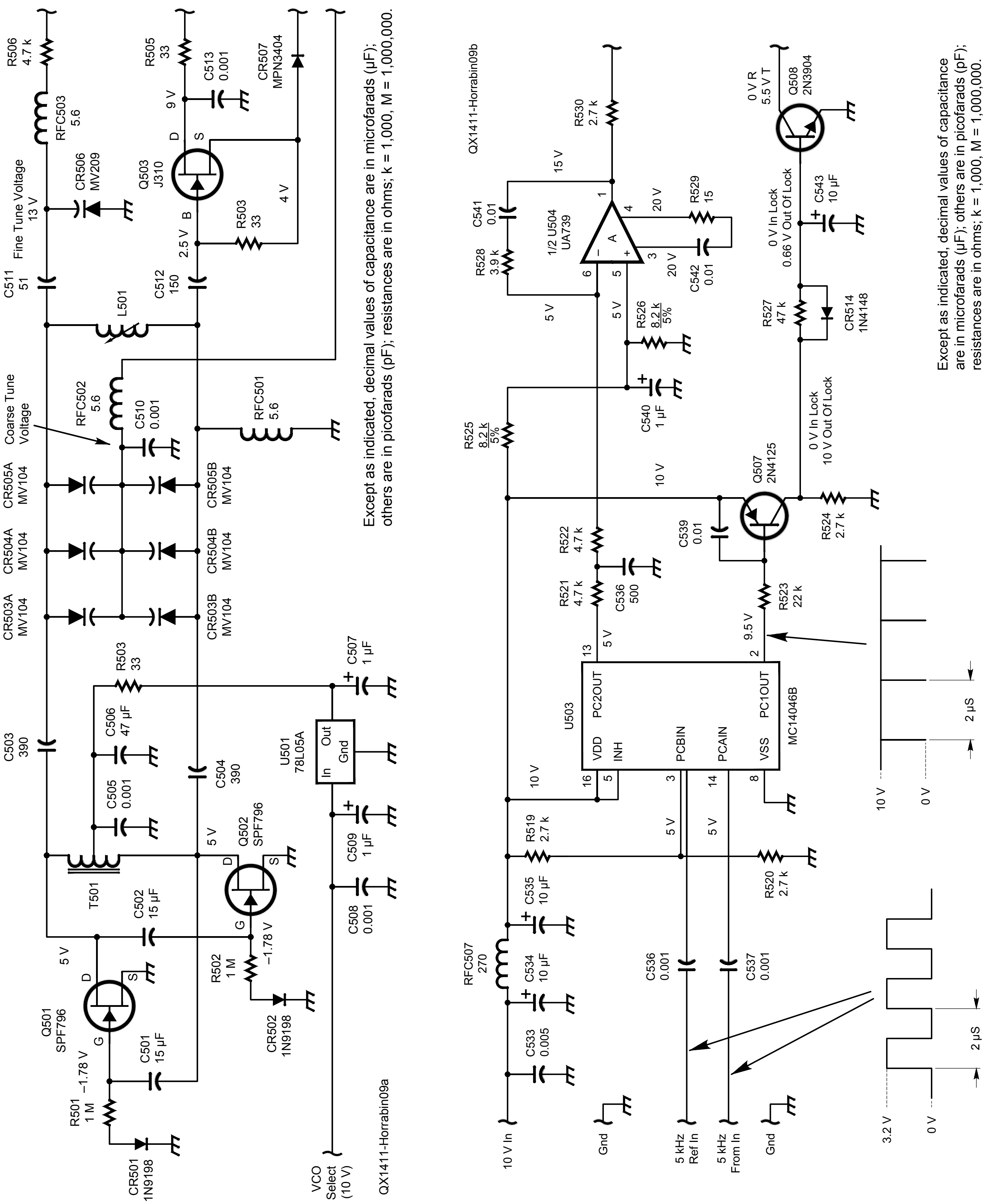

Figure 9 — Part A shows a portion of the Drake TR7 VCO. This circuit is very similar to the G3PDM design. In looking at the 78L05A voltage regulator, the 33 Ω resistor (R503) and 47 µF capacitor (C506) would not be effective at suppressing close-in noise. Part B shows a portion of the VCO amplifier circuit. Note that there is no capacitor across the 3.9 kΩ resistor (R528), which would have helped roll off the op amp gain above the PLL bandwidth.

sponsor. I will always be grateful to BAC for the opportunity that they gave me and the experience gained by working with such a great and talented bunch of people.

Peter Mcintosh, who was on Daresbury Laboratory's Young Scientist Apprenticeship scheme gave me a hand with some circuit board design work for the test oscillators and has now moved on to greater things. After spending some years working in the USA, Peter came back to Daresbury Laboratory and is now head of ASTEC (Accelerator Science and technology).

The ideas of Wes Hayward, W7ZOI, must be singled out yet again for particular praise. He is not known as "The King" for nothing. Along with Doug DeMaw, W1FB, (SK) he wrote the book that became my "bible;" *Solid State Design for the Radio Amateur*. There is something that is typical of the best in American RF engineering ideas about this book and also something that is timeless about it. I still pick it up to read it even though it's a bit dog eared by now.

I would also like to acknowledge Spectrum Software and their circuit simulator, *Micro-Cap 3*. Without this valuable tool, I could have spent years trying different circuit configurations and never have gotten a result.

Thanks also to my friend Martein Bakker, PA3AKE, for doing the phase slope simulations of the double tank and G3PDM oscillators using *Spice*, and for the phase noise measurements of the local oscillator on his Drake TR7 transceiver.

I must also thank the Radio Society of Great Britain for their permission to show the circuit of the G3PDM VCO from the *Radio Communication Handbook*.

Finally, an unsung hero until now is my former second in command, Andrew Moss, who is now Senior RF Engineer at Daresbury Laboratory. He gave me invaluable help with the experimental work that was done for the HF7070 receiver front end. Not only that, low phase noise (and low time jitter) oscillators is the name of the game these days for particle accelerators. His support for the work that Martein, PA3AKE, did with phase noise on the AD9910 DDS chip looks like it is paying off in the application of the Analog Devices 3.5 GHz DDS chip, the AD9914, in state of the art RF systems under development at Daresbury Lab.

Useful Reading Material

Colin Horrabin, G3SBI, "The HF7070 HF/LF Communications Receiver Prototype," Jul/Aug 2013 *QEX*, pp 37-44.

Pat Hawker, G3VA, "G3SBI's High Performance Mixer (H-Mode Mixer)," Technical Topics, *RadCom,* Oct 1993 pp 55-56.

Pat Hawker, G3VA, "Crystal Filters for High Performance Mixers" (H-Mode Mixer termination with crystal filters using quadrature hybrids), Technical Topics, *RadCom,* Jan 1994 pp 37-39.

Jacob Makhinson, N6NWP. "A Termination Insensitive Amplifier," *QEX*, July 1995, pp 21-29.

Jacob Makhinson, N6NWP, "A Drift Free VFO," *QST*, Dec 1996, pp 32-36.

"The Frequency Synthesizer on the G3PDM receiver," *The Radio Communication Handbook* (RSGB) fifth edition 1982 p 4-53.

Pat Hawker, G3VA, "Towards The Superlinear Receiver" (low noise double tank oscillators), Technical Topics, *RadCom*, July 1994, pp 53-55.

"The Double Tank Oscillator VCO Circuit for the AR7030 Receiver." You can find this on the Warrington Amateur Radio Club website (**http://warc.org.uk/**), under the Projects tab, in the CDG2000 Transceiver project section.

Wes Hayward, W7ZOI, and Doug DeMaw, W1FB, (SK) *Solid State Design for the Radio Amateur*, ARRL, 1986 (Out of print. You may find a copy with an old-time ham or a school or technical library.)

Wes Hayward, W7ZOI, Rick Campbell, KK7B, and Bob Larkin, W7PUA, *Experimental Methods in RF Design*, ARRL, 2003, ISBN: 0-87259-879-9, ARRL Order No. 8799, $49.99. ARRL publications are available from your local ARRL dealer or from the ARRL Bookstore. Telephone toll free in the US: 888-277-5289, or call 860-594-0355, fax 860-594-0303; **www.arrl.org/shop**; **pubsales@arrl.org**.

William Sabin and Edgar Schoenike, *Single Sidebands Systems and Circuits*, 2nd Edition, McGraw Hill, 1995.

Notes

[1] Martein Bakker, PA3AKE, This URL is for Martein's discussion of his Sideband Noise Measurement methods: **http://martein.home.xs4all.nl/pa3ake/hmode/dds_pmnoise_intro.html**.

[2] Wes Hayward, W7ZOI, "Oscillator Noise Evaluation with a Crystal Notch Filter, Jul/Aug 2008 *QEX*, pp 6-12.

[3] D. B. Leeson, "A Simple Model of Feedback Oscillator Noise Spectrum," *Proceedings of the IEEE*, February 1966.

[4] Randall W. Rhea, *Oscillator Design & Computer Simulation* (Second ed.), McGraw-Hill, 1997, ISBN 0-07-052415-7, p 115.

Colin Horrabin, G3SBI, was born in 1941. His father provided him with a World War II BC348 radio receiver for his 12th birthday, followed by a copy of the ARRL Handbook *for Christmas. After years building various projects using government surplus equipment, he obtained his Amateur Radio license in 1963. He has a degree in electrical engineering and a degree equivalent qualification in mechanical engineering. Following an apprenticeship with the British Aircraft Corporation in the early 1960s, he spent over 30 years working at Daresbury Laboratory as an electronic engineer. Colin is interested in small DX antennas for the LF bands, and intends to do some work on small multi turn spiral wound loops that are self resonant containing ¼ wavelengths of wire, which are suitable for transmitting.*

Robert J. Zavrel, Jr, W7SX, ARRL Technical Advisor

PO Box 91, Elmira, OR 97437; **w7sx@arrl.net**

Bob Zepp: A Low Band, Low Cost, High Performance Antenna

This antenna array provides a switchable, 4-direction, vertically polarized, full-azimuth-coverage high gain antenna for 160 meters and a bidirectional horizontal antenna for 40/75/80 meters. In Part 1. Bob takes us through the early development stages of his antenna.

In the Feb 2010 issue of *QST* I presented "The Curtain-Zepp."[1] That wire antenna array was supported by two 140 foot Douglas Fir trees, which provided bidirectional gain on the three low bands: 160, 80/75 and 40 meters. It has proven to be a very effective antenna on all these bands. This paper presents a revised version of the Curtain Zepp. In addition, several steps leading to the final design are presented, therefore this paper describes several possible low-band antenna configurations, from the simplest to the more complex Bob-Zepp. An antenna builder could also use these steps as a progression to more complex and more effective arrays as confidence is gained at each step. This, in effect, is what I did in the development of this array over an eight year period.

The Bob-Zepp is also scalable for higher frequencies. For example, instead of 160/80/40 meters, it could be configured as an 80/40/20 meter or even a 40/20/10 meter array, the latter would be about 30 feet high and 40 feet long. *EZNEC* or a similar design tool would be required to optimize such alternate designs. Also, the dimensions are not critical. The mini Bobtail configuration "wants to work." Optimization should be done using modeling tools.

Advantages of the final version:

1) Bidirectional horizontal gain (dBd) for 80/75/40 meters.

[1]Notes appear on page 48.

Photo A — Here is the antenna feed point. The ladder line from the shack (not visible in this photo) comes to the pole from the left, then down to the tuner box. The coax and control cables are fed though buried electrical conduit. You can see a corner of my outdoor patio, where the tuner box serves as a convenient summer serving table.

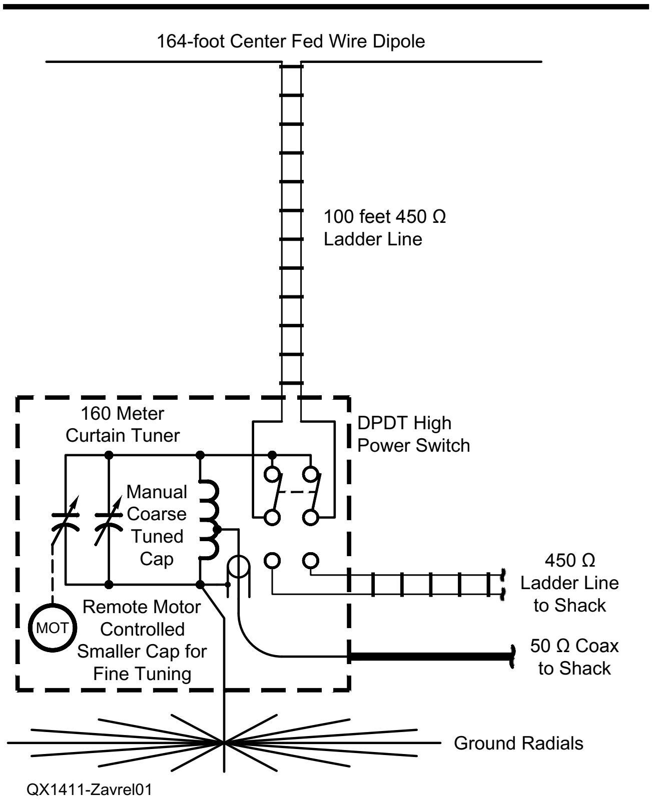

Figure 1 — Step 1: An extended double Zepp on 40 meters, a dipole on 75/80 meters, and top-loaded vertical for 160 meters. and three plots showing azimuth vs. gain on the three bands.

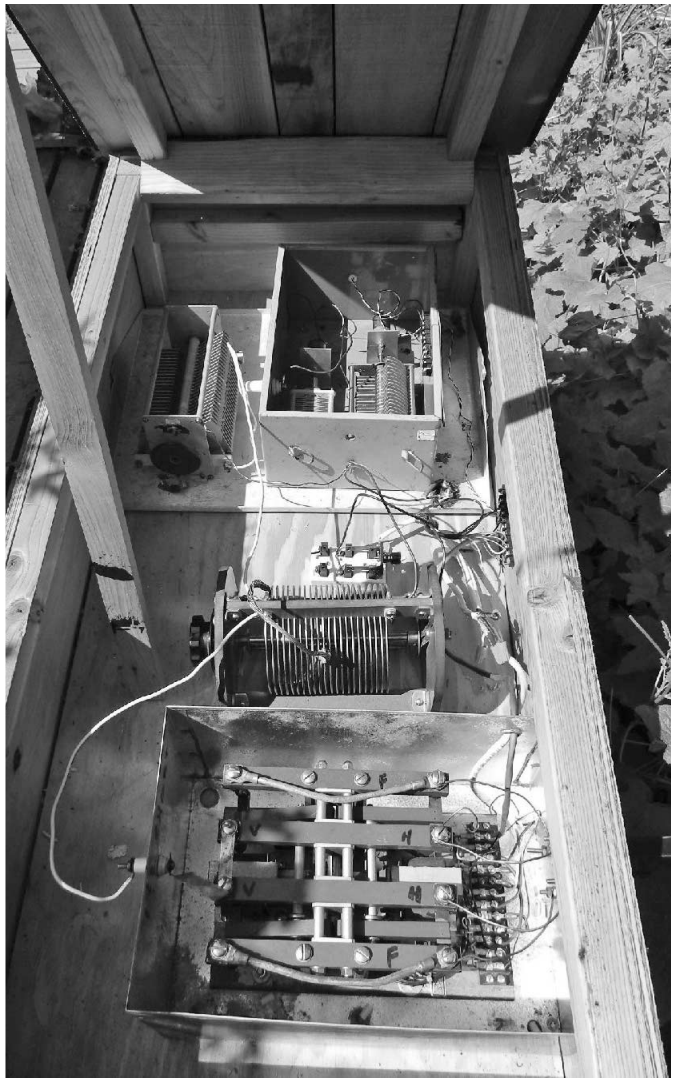

Photo B — This photo is a view of the tuning networks inside the tuning box from Photo A. Near the bottom of the photo you can see my heavy duty antenna transfer switch. This was surplus from a high power AM broadcast facility. Vacuum switches or high voltage ceramic switches can also be used.

2) Vertically-polarized, switchable all-azimuth coverage gain antenna for 160 meters.

3) Minimal ground radial system (6 short wires only on the fed element).

4) Two or optional four-support system.

5) Inexpensive wire construction that can be supported by trees.

6) All tuning and switching performed at near-ground level and with remote control from the shack.

The final version also includes some techniques that may be useful for a wide variety of antenna applications outside of the Bob-Zepp. For example:

1) Gain optimization of a horizontal antenna using "drooping" vertical end loading.

2) Suggestions on dressing open wire feed-lines that feed wire antennas prone to wind movement (tree supported).

3) Developing an end-fire pattern from a Bobtail antenna, first bidirectional, then switchable mono-directional.

4) Considerable information on using quad-style loop parasitic elements with ground-mounted verticals.

5) An alternative method for implementing a remote control motor-driven variable inductor.

The First Three Steps

There are two main differences between the original and this antenna: horizontal polarization on 80/75 meters and a four-directional vertical polarized capability on 160 meters. On 80 meters the low elevation angle gain with the old vertical curtain array was comparable with a more optimized horizontal configuration. Furthermore, band-switching and tuning involved a rather complex circuit for vertical polarization on 80/75 meters. After extensive modeling with *EZNEC* and lots of experimentation, I settled on the addition of two end-fire configurations that provide mono-directional "east" and "west" patterns for 160 meters.

Switching back to the old curtain configuration, a third north-south bidirectional pattern emerges, providing full-azimuth coverage from the one single antenna (not to mention an excellent bidirectional horizontal antenna for 80/75 and 40 meters)! So this new array can be considered a vertical polarized directional array with full azimuth coverage on 160 meters and a high gain bidirectional horizontal polarized array for 80/75 and 40 meters. This array is supported by the same two trees and is the same size as the Curtain-Zepp.

In addition, a reflecting loop in back of the broadside curtain, supported by two other trees forms a unidirectional 160 meter array with substantial gain toward Europe, the Middle East and East Africa, all very difficult paths from Oregon on the low bands. This loop can be detuned remotely to rees-

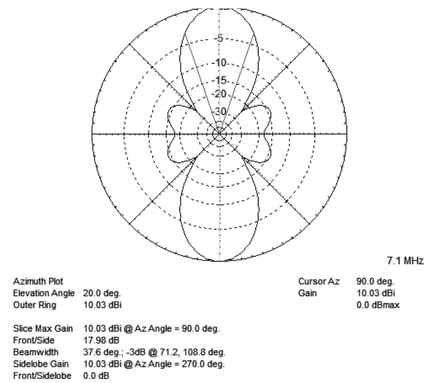

Figure 2 — This azimuth versus gain *EZNEC* pattern plot shows the 40 meter dipole.

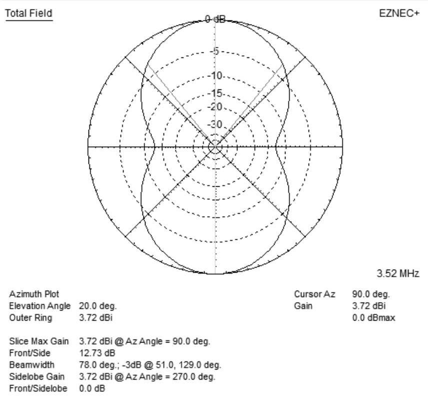

Figure 3 — This azimuth versus gain *EZNEC* pattern plot shows the 80 meter dipole.

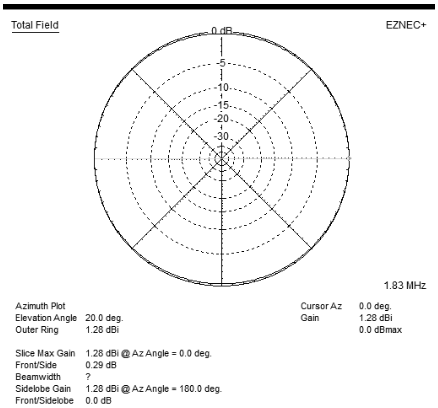

Figure 4 — Here is the azimuth versus gain *EZNEC* pattern plot for the 160 meter top loaded vertical.

Photo C — Here is my preferred method of vertical ladder line mounting. Coming down from the antenna feed point, the ladder line is fed through a plastic electric fence insulator (at the top right corner of the photo) to allow it to move up and down with the wind. Placing this guide for the line about 8 feet high prevents the line from becoming entangled with garden plants and people. Near the bottom of the photo, a ¼ inch nylon nut and bolt fasten the line to a rubber bungee cord, which is then fastened to a heavy weight on the ground. The rest of the secured line is then fed to the tuning box. Note that this photo shows an older version of the tuning box on the left. Photos A and B show the current tuning box.

tablish a bidirectional pattern for coverage of the south-west Pacific. So on 160 meters, four separate mono-directional patterns are available. The loop was intentionally made smaller than a full size 160 meter reflecting loop. Two inductors are included in the loop to optimize its functions as a director or reflector. There will be more on optimizing loops in Part 2, in the Optimizing Loop Parasitic Elements section.

Note: I use azimuth directions north, south east and west in this paper for simplicity. Of course this array can be oriented in any configuration. My "north" is actually at 30° azimuth, directly toward Europe from my QTH. Therefore, "east" is 120°, south is 210°, and west is 300°.

Step 1:

Tree supported extended double Zepp antennas have been my default favorite antenna on 40 meters for over 20 years.[2] Therefore this was my starting point: a 164 foot wire, center fed with 450 Ω ladder line, strung at a height of about 100 feet between two trees. This also forms a very effective "long dipole" on 75/80 meters. The ladder line is fed to a rack in my shack, where tuners achieve a perfect match anywhere in the 80/75/40 meter bands. On these bands it forms a north-south bidirectional pattern. It can also be fed as a short dipole on 160 meters. A 160 meter vertical can be created by breaking the ladder line near ground level, shorting the ladder line wires, and feed it against the ground as a familiar "T" antenna. This requires a second transmission line (coax) fed from the shack, and of course the required tuning and switch-

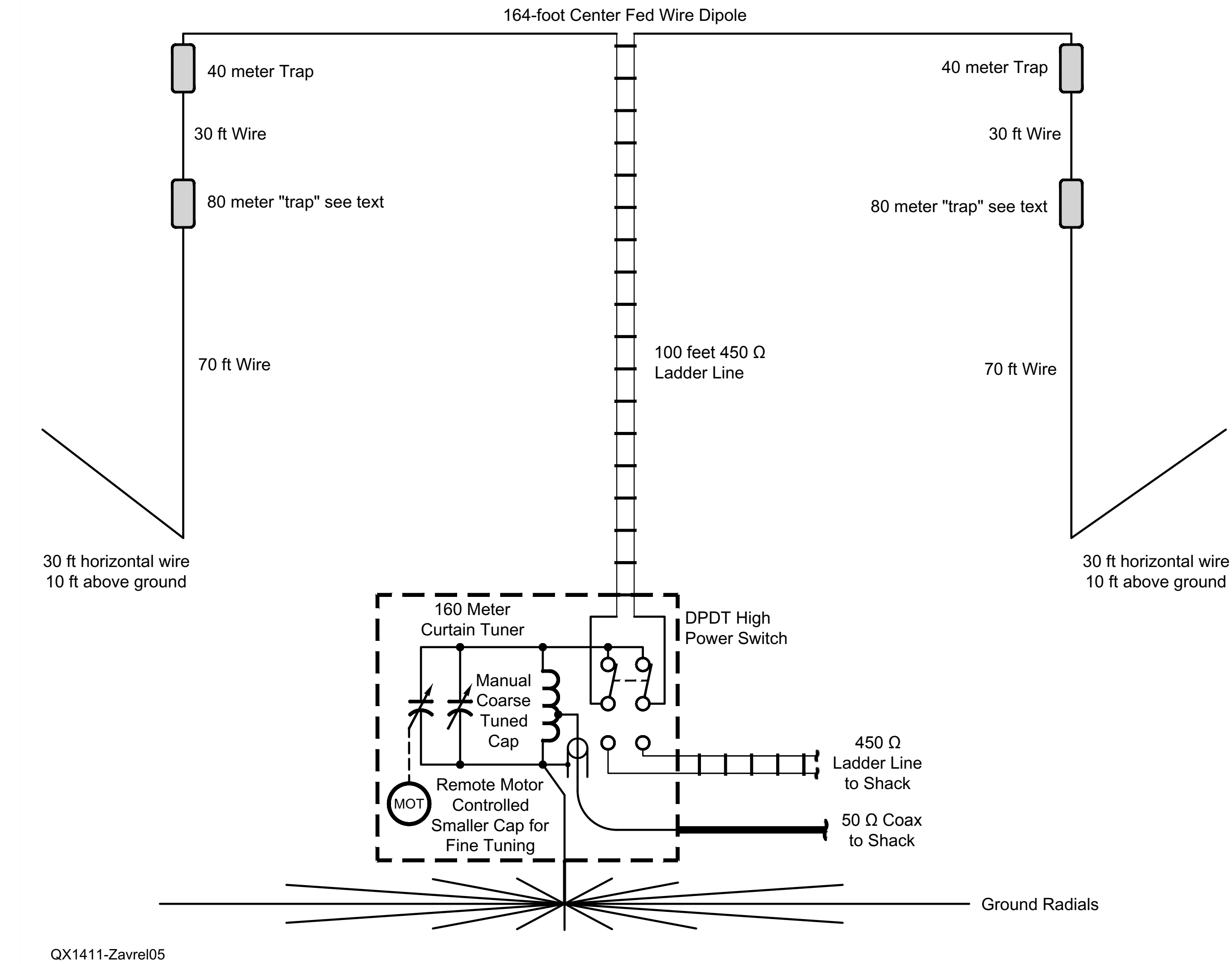

Figure 5 — Step 2: An extended double Zepp on 40 meters, extended dipole on 75/80 meters and a bidirectional curtain for 160 meters.

ing circuitry at the base. This antenna is very effective on 160 meters as a single-element monopole. On 160 meters, the horizontal wires form a capacitive hat, thus moving the current maximum away from the base to near the top of the vertical element (shorted ladder line). This increase in base-fed impedance lowers the effect of ground loss on efficiency and thus this vertical needs only a modest ground radial system. I use only four 50 foot radials and a 6 foot copper ground rod. Figure 1 illustrates this arrangement. Figures 2, 3 and 4 show the *EZNEC* patterns for the 40 and 80 meter horizontal configurations and the 160 meter vertical.

Photo A shows the antenna feed point. The ladder line from the shack (not visible in this photo) comes to the pole from the left, then down to the tuner box. The coax and control cables are fed though buried electrical conduit. This tuner box doubles as a food serving table for summer outdoor activities, when the low bands are dead! It is *imperative* that all such tuning points be secured from people and animals when the array is in use!

Photo B is a view of the tuning networks inside the tuning box. Near the bottom of the photo you can see the antenna transfer switch. This was surplus from a high power AM broadcast facility. Vacuum switches or high voltage ceramic switches can also be used.

Photo C shows my preferred method of vertical ladder line mounting. Coming down from the antenna feed point, the ladder line is fed through a plastic electric fence insulator (at the top right corner of the photo) to allow it to move up and down with the wind. Placing this guide for the line about 8 feet high prevents the line from becoming entangled with garden plants and people. A ¼ inch nylon nut and bolt fasten the line to a rubber bungee cord, which is fastened to a heavy weight on the ground. The rest of the secured line is then fed to the tuning box, which houses the heavy transfer switch. Ladder line with braided wire is preferred for this section of the line rather than solid copper wire.

Step 2

I want vertical gain on 160 meters! If we drop vertical wires from the horizontal wire ends, an array that resembles a Bobtail Curtain emerges. In addition traps, for 40 and 80/75 meters are also included to maintain the extended double Zepp response on 40 meters and optimize broadside gain on 80/75 meters. My original array was configured as a curtain for both 160 and 80/75 meters. Modeling and on-air testing both indicated that returning to horizontal polarization on 80/75 meters would be to my advantage. To use horizontal polarization on 80/75 meters, a second set of traps are required, since the vertical wires provide an undesirable pattern on that band. This

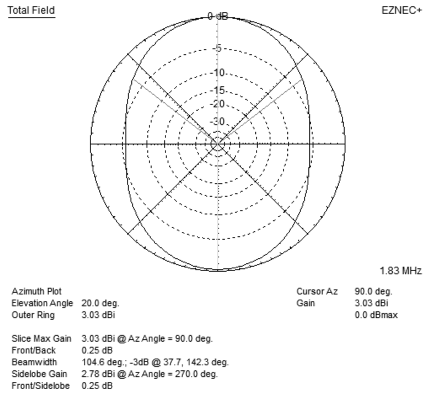

Figure 6 — This plot shows the 160 meter bidirectional curtain north/south pattern.

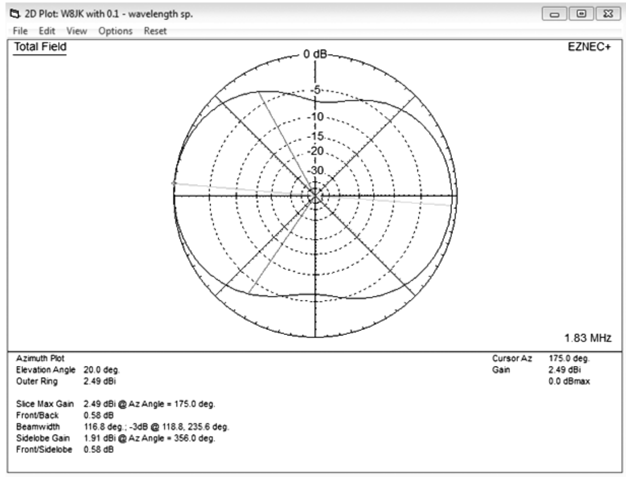

Figure 7 — Here is the plot showing the east/west pattern formed when the new array is fed with the ladder line, identical to the feed method on 40/80/75 meters, but the result is a two-element end-fire bidirectional vertical array.

modification also greatly simplified the tuning and switching requirements at the base of the array, freeing expensive components that would prove far more useful in the more complex arrays that follow.

By moving the 75/80 meter traps "down" the end vertical elements, the broadside gain increases over a simple 164 foot dipole, until reaching a point where further lowering of the traps begins reducing the gain. This point is about 30 feet below the tops of the vertical wires, the "sweet spot" for 80/75 meters. The 80 meter traps use a rather high C/L ratio since higher inductance proved to be a disadvantage for all patterns on 160 meters. The 80 meter traps use 200 pF capacitors and 9.5 μH inductors. The 40 meter traps use 40 pF and 13 μH components. I use high voltage doorknob capacitors (at least 10 kV), and wind #10 insulated copper wire on ABS pipe sections with the capacitors placed inside the pipe and a pipe cap on top to keep the inside of the pipe (and capacitors) dry. Also, I drilled small drainage holes in the bottom of the ABS to avoid condensation inside the traps.

Finally, for 160 meters, low horizontal "capacitive boots" are placed at the bottoms of the vertical wires. Since the vertical end wires are effectively top fed vertical elements, the loading lines perform the identical function of capacitive hats in base fed verticals. That is why I call them "capacitive boots." These have the effect of optimum placement of the RF current along the array elements for maximum gain of the array. This completes the tri-band bidirectional array. The result is a north/south bidirectional array with more gain on 80/75/160 meters.

As mentioned before, the 160 meter curtain is configured by shorting the ladder line leads together at the base and feeding it as a vertical against the ground, resembling a full-sized Bobtail Curtain. On 160 meters, however, if we feed this array with the balanced ladder line at the center of the horizontal wire (again, like a dipole), the array dimensions are conducive to forming a two-element vertical end-fire array (east/west) on 160 meters! Thus, on 160 meters we now have two orthogonal bidirectional patterns for full-azimuth coverage *in addition* to the Bobtail bidirectional pattern. Of course, the end-fire configuration is also tuned in the shack for a perfect match. The result is that we can switch between two bidirectional 160 meter patterns (north/south or east/west) simply by changing the way we feed the array! Figure 5 illustrates this construction.

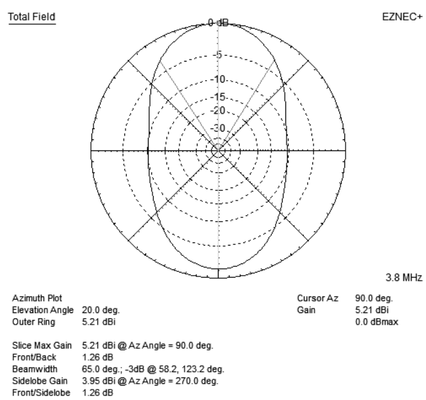

Figure 8 — This plot shows the higher broadside gain realized on 75 meters, resulting from optimum placement of the 80/75 meter traps. The 40 meter response is essentially identical to the plot shown in Figure 7.

Figure 6 shows the 160 meter bidirection curtain north/south pattern. Figure 7 shows the east/west pattern formed when the new array is fed with the ladder line, identical to the feed method on 40/80/75 meters, but the result is a two-element end-fire bidirectional vertical array. Finally, Figure 8 shows the higher broadside gain realized on 75 meters resulting from optimum placement of the 80/75 meter traps. The 40 meter response is essentially identical to the pattern shown in Figure 7.

In Part 2 of this article, I will describe how the end-fire radiation pattern can become a switchable dual mono-end-fed directional array. This will increase the gain in these two directions. I will also describe the remaining steps to design the complete Bob-Zepp antenna.

Bob Zavrel, W7SX, is an ARRL Life Member, Technical Advisor and Amateur Extra class licensee. He has been licensed since 1966. His primary interest in Amateur Radio is low band DXing and designing and building antennas, tuners, and amplifiers. Bob holds 5BDXCC, 5BWAZ (200), has 334 mixed, and 324 CW entities confirmed. Previous call signs include WN9RAT, WA9RAT, WA9RAT/HR2 and SV1/W7SX.

Bob has a BS in Physics from the University of Oregon and has worked in RF engineering for over 30 years. He has five patents, and has published over 50 papers in professional and Amateur Radio publications, including the first block diagram of an SDR receiver in 1987. He was involved with the first generation of RF integrated circuits for cellular phones, and worked extensively with DDS, WLAN and passive mixer development. Bob is currently an RF Research and Development Engineer for Trimble Navigation with a primary focus on high precision GPS, down to mm accuracy.

Notes

[1]Bob Zavrel, W7SX, "The Curtain Zepp — A Bidirectional Antenna for 160, 80 and 40 Meters," Feb 2010 *QST*, pp 36 – 39.

[2]Bob Zavrel, W7SX, "Maximizing Radiation Resistance in Vertical Antennas," Jul/Aug 2009 *QEX*, pp 28 – 33.