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March/April 2020

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A Forum for Communications Experimenters

Issue No. 319

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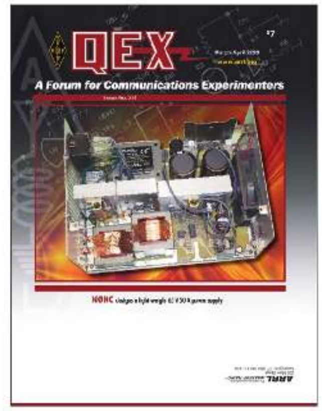
March/April 2020

About the Cover

Ralph J. Crumrine, NØKC, designed a light-weight 3 kW power supply to support amplifiers based on the newer 65 V LDMOS FET devices. The power supply uses state-of-the-art switching MOSFETS supplying a 65 V output at up to 50 A, and includes some features particularly useful to the amateur radio operator/experimenter. The design uses a pulse width modulated waveform operating at 60 kHz, and user-wound ferrite transformers. The resultant new design is housed in a 5 1/2" H by 7-1/2" W by 10" D cabinet and weighs just 9 lb.



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Kazimierz "Kai" Siwiak, KE4PT

Perspectives

QEX Online Edition and More Morse Challenge Results

Very soon *QEX* will be available online as an ARRL member benefit. The printed edition of *QEX* will still be available to all subscribers including non-members. *QEX* Online will be the same as the printed magazine, but will feature full color. It joins the ARRL suite of digital magazines — *QST* Online, *On the Air*, and *NCJ* Online — as member benefits. For *QEX* authors that means a greatly expanded audience!

The *Morse Input Design Challenge* ended in December 2019, with gratifying results. In the previous *Perspectives* we announced that the earliest complete entry was received from Joseph M. Haas, KEØFF. Here we can announce that entries were received from six additional participants of the *Challenge*. They are Jonathan A. Titus, KZ1G; Gwendolyn S. Patton, NG3P; Ralph L. Irons, N4RLI; Robert D. Anding, AA5OY; Kevin A. Ames, AA7YQ; and David Bern, W2LNX. Each will receive a free one-year subscription (or subscription extension) to the printed edition of *QEX*. You can expect articles describing the entries to appear soon.

In This Issue

Joseph M. Haas, KEØFF, adds DTMF capability to popular transceiver microphones.

Jacek Pawlowski, SP3L, describes a wide range of new broadband wire antennas in this Part 1 of a longer article.

Ralph J. Crumrine, NØKC, designs a light weight 3 kW power supply to support state-of-the-art switching MOSFET amplifiers.

Eric P. Nichols, KL7AJ, in this first-of-a-series, reveals the simplicity and many uses of the double balanced mixer.

Michael P. Hasselbeck's, WB2FKO, *Technical Note* describes anchors for a 3D printed lid.

Kai Siwiak's, KE4PT, *Technical Note* explains why typical small HF loop antennas are not magnetic loops.

Writing for *QEX*

Keep the full-length *QEX* articles flowing in, or share a **Technical Note** of several hundred words in length plus a figure or two. Let us know that your submission is intended as a **Note**. *QEX* is edited by Kazimierz "Kai" Siwiak, KE4PT, (**ksiwiak@arrl.org**) and is published bimonthly. *QEX* is a forum for the free exchange of ideas among communications experimenters. The content is driven by you, the reader and prospective author. The subscription rate (six issues per year) in the United States is $29. First class delivery in the US is available at an annual rate of $40. For international subscribers, including those in Canada and Mexico, *QEX* can be delivered by airmail for $35 annually. Subscribe today at **www.arrl.org/qex**.

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Very best regards,

Kazimierz "Kai" Siwiak, KE4PT

Eric P. Nichols, KL7AJ

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The Versatile Double Balanced Mixer

This first-in-a-series reveals the simplicity of the double balanced mixer and its many uses.

The double balanced mixer (DBM) is a nearly ubiquitous component in nearly every amateur radio transmitter and receiver. The DBM most commonly appears as a monolithic, modular building block, around which a number of interesting radio circuits can be assembled. While the concept of the balanced mixer goes back to the vacuum tube days, it wasn't until high performance diodes such as the Schottky diode and the wide bandwidth ferrite core transformer became readily available that the use of the modern DBM took off.

A great deal of amateur and commercial radio literature has explored the use and optimization of the DBM and associated circuitry as frequency converters, especially with regard to dynamic range. Since this topic has been so thoroughly covered elsewhere, we will not go into this aspect in great detail in this series. Instead, we will turn our attention, in the next few articles, on the more unheralded applications of this versatile building block.

The modern, modular DBM is well at home in a number of roles, such as a modulator, demodulator, phase shifter, phase detector, linear attenuator, RF switch, and even an audio special effects generator.

A Bit of History

One of the earliest references to the present manifestation of the DBM appeared in the April 1969 *QST* article, "Some Notes on the Solid-State Product Detector" by Doug DeMaw, W1CER. In this article, four types of product detectors are described; the second example is the present day DBM which we will explore. This circuit originally had several different names, one of which was "ring modulator," because four diodes are arranged in a ring configuration. In the original drawing, (**Figure 1**) it isn't quite as obvious that the diodes are in a ring until you reorient the components slightly (**Figure 2**). I first heard the term ring modulator in reference to the Moog Synthesizer, as it was responsible for a number of heretofore unheard of audio effects that the Moog produced. We'll explore this a bit more in our later installment, "Fun and Games with the DBM." The terminology was eventually cast as the DBM, for the most part.

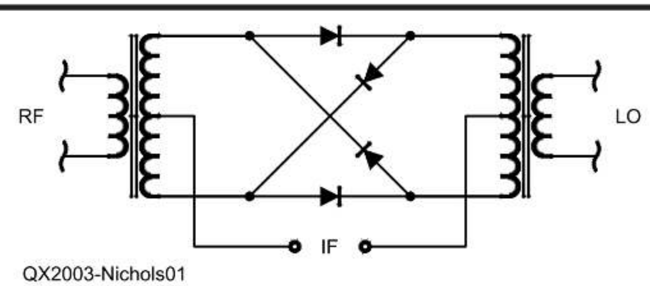


Figure 1 — The double-balanced mixer — sometimes called the doubly-balanced mixer — is a deceptively simple circuit, with numerous interesting and useful applications.

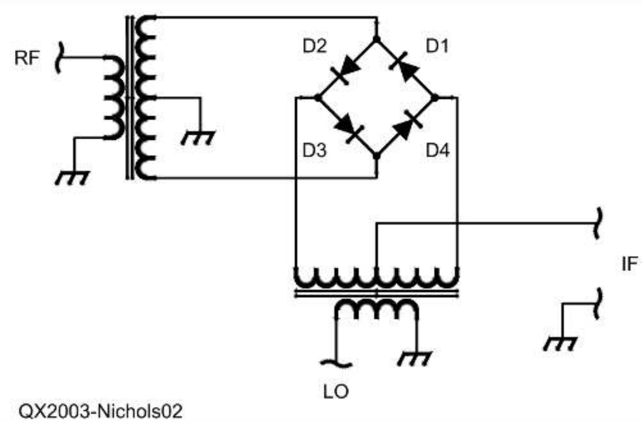


Figure 2 — It's sometimes helpful to rearrange components of identical circuits to get a better grasp of their function. This ring modulator is identical to that of Figure 1.

Passive With a Purpose

While a balanced modulator can be built with amplifying devices, including vacuum tubes — as mentioned earlier — the standard modern DBM uses passive diodes. By some definitions, a diode is an active device, so we won't need to belabor the point. However, the important factor is that the diode DBM doesn't amplify anything; in fact it imparts a net loss to the signals it processes. But the DBM makes up for this shortcoming in both simplicity and in agility.

What it Is

Regardless of the actual specific role the DBM plays, its function is always the same; it multiplies one signal by another. We can infer this because of its original role as a product detector, a device that gives us an audio signal which is the instantaneous product of the beat frequency oscillator (BFO) voltage and an IF voltage of a receiver. As we look under the hood of the DBM, we will discover exactly how it does this; it is intuitive and yes, somehow surprising.

When viewed as a ring modulator, we can see that the circuit bears a certain resemblance to the Wheatstone bridge. Indeed it is a form of bridge. It's also fairly evident as to why it's called a ring modulator, as there are four diodes, all pointing in the same direction around a ring. For the time being, we can consider the diodes as ideal one-way switches; we'll discuss the non-linear aspects of the diodes later.

If we apply RF to the 'top' and 'bottom' of the ring (Figure 2) via the RF input transformer, we find that D2 and D3 conduct during the positive half of the input cycle, and D1 and D4 conduct during the negative half of the cycle. Neglecting the forward voltage drop of the diodes, we see that the input signal is effectively applied across a dead short. At no point in time do we see a difference of potential between the horizontally opposed corners of the ring. No signal voltage appears at the LO transformer. In its resting state, the DBM is perfectly balanced; no signal is transferred from the RF to the LO port, or vice versa.

The third port, or IF port appears between the center taps of the two transformers. In this example, the center tap is grounded on the RF port, but both center taps can be left floating if necessary.

Now, since there is a dc path from the IF port to the diodes, we can upset the balance of the diodes by applying a dc voltage to the IF port. Note that while tradition tells us that the IF port should be an *output* signal — such as when used in a receiver — the IF port can be an input as well, and it is in this mode that some of its most interesting properties are demonstrated. Perhaps the most straightforward demonstration of this is when the device is used as an RF switch. If we apply a positive voltage to the IF port, we will find that D3 and D1 are in conduction, while D2 and D4 are reverse biased. There is now an RF path from the bottom of the RF input transformer to the left side of the LO transformer. There is also a path from the top of the RF input transformer to the right side of the LO transformer. Therefore an RF signal can pass essentially unattenuated between the RF and the LO ports.

If we apply a *negative* potential to the IF port, we find that D4 and D2 conduct, while D1 and D3 are reverse biased. The signal path is now from the top side of the input transformer to the left side of the LO transformer, and from the bottom of the input transformer to the right side of the LO transformer. Once again, the signal passes nearly unattenuated from the RF to the LO port, but in the opposite phase. As you might imagine, this can be a very useful property. Perhaps the most obvious is the ability to create bipolar phase shift keying by simply applying a positive (mark), and a negative (space) voltage to the IF port.

To demonstrate this polarity reversal using my homebrew DBM (**Figure 3**) I use an asymmetrical RF waveform for easy identification. The upper trace is the input signal and the lower trace is the output signal. **Figure 4** shows the signals in phase,

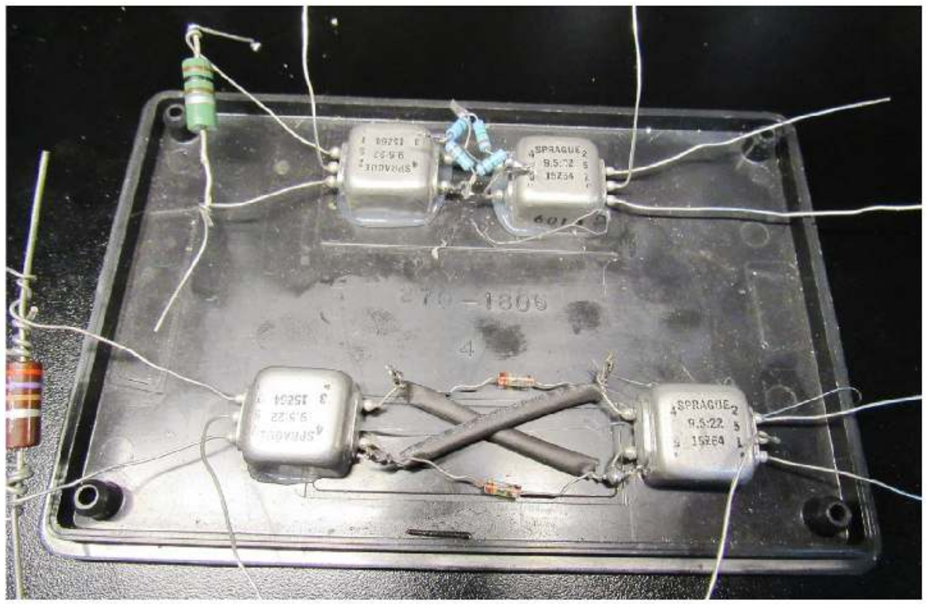


Figure 3 — First you make it work; then you make it pretty. Four 'mystery' junk box IF transformers are employed to create the DBMs used to generate the waveforms in this article.

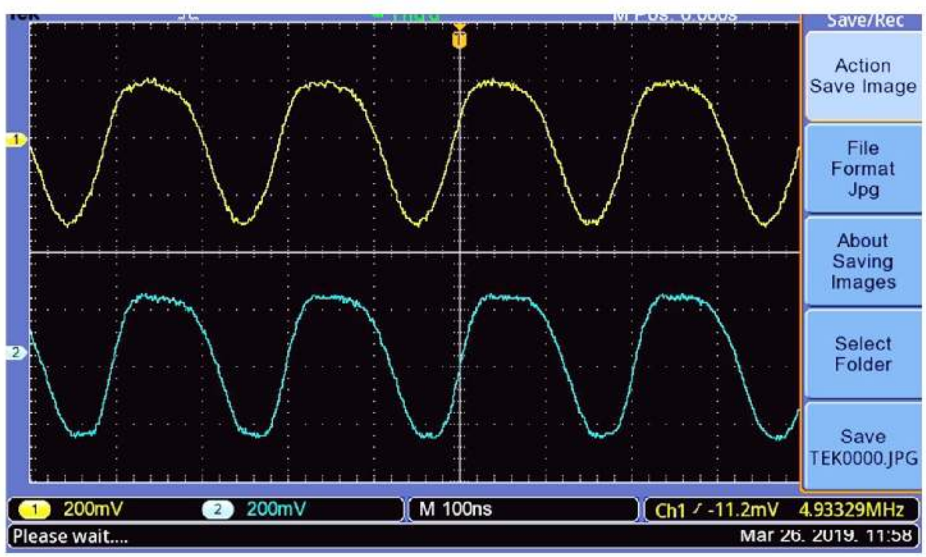


Figure 4 — A slightly 'dented' RF waveform is used to readily identify the relative phasing of the DBM input (upper trace) and output (lower trace) signal. In this display, there is no phase shift through the DBM.

using a positive voltage on the IF port, and **Figure 5** shows the signals out of phase, using a negative voltage on the IF port.

Hard Driven

When operating as a switch, it's important that the diodes are driven 'hard' by the dc voltage to assure both a minimum attenuation as well as minimum distortion of the RF waveform passing through. The dc voltage should be many times the value of the RF voltage, so that there is no danger of the diode failing to conduct during any portion of the RF cycle. When driven 'hard' even a mediocre diode can closely simulate a piece of wire, and is an effective low distortion switch.

Attenuator

If somewhat less dc voltage is applied to the IF port, the DBM can also serve as an RF attenuator. The signal paths are the same as in the above-described bipolar switch, but by reducing the dc drive, one can incur some signal loss between the RF and LO ports. This is because when a diode is operated in the square law region. It actually serves as a multiplier, where the current passing through the diode is the product of the applied voltages, in this case one being an RF voltage and one being a dc voltage. When used as an attenuator, it is actually preferable to use rather high RF signal levels and a low dc voltage, as this results in the most linear control of the attenuation as a function of dc voltage. One thing you want to avoid is running the RF signal voltages and dc voltages at approximately the same level, as this will result in the maximum amount of distortion.

Not So Square Law

It should be noted at this point that the so-called square law region of a diode is a bit of an over-simplification. When a diode is forward biased just above the conduction region — about 0.6 V for a silicon device — the current *vs.* voltage curve is not actually a square function, but rather has a continuously varying exponential base! However, these various regions are very small, and on the average, the 'curved' regime of a diode current *vs.* voltage curve is approximately a square-law function. Because the diode curve transits a continuum of exponential bases, it can be used as the core of a logarithmic or exponential amplifier. However, the design of such a device requires great care. The general 'square law' region, however, is wide enough that it is easy to use for multiplication purposes. The Shockley diode equation (**https://en.wikipedia.org/wiki/Shockley_diode_equation**) describes this exponential function in detail.

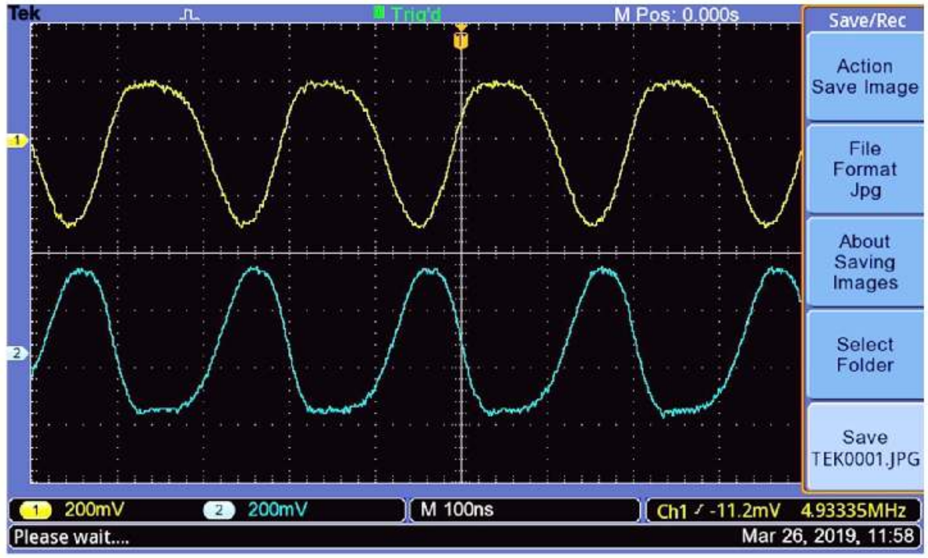


Figure 5 — Here you can see the inverted waveform (lower trace) emerging from the DBM when a negative voltage applied to the IF — or modulation — port.

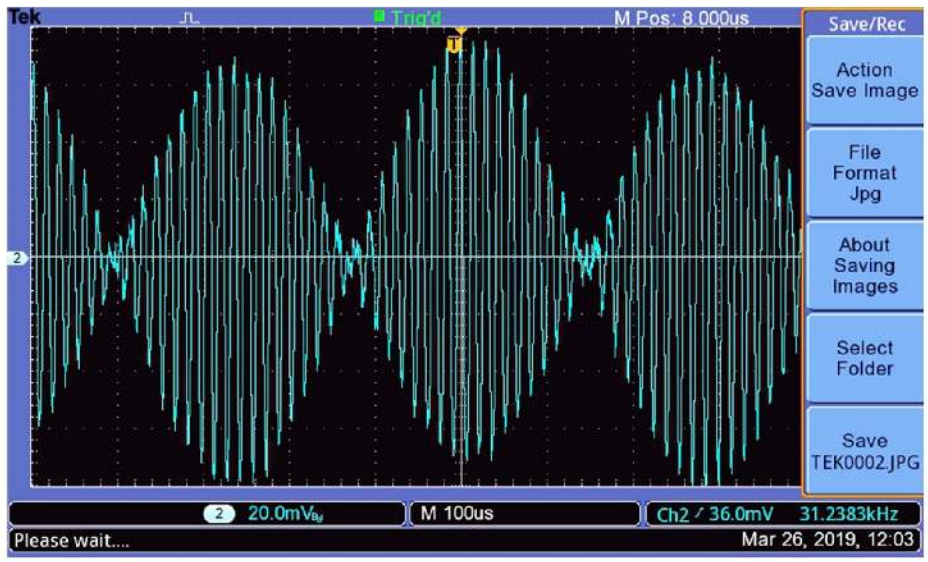


Figure 6 — A double sideband suppressed carrier signal is easily generated with a DBM.

A Versatile Modulator

In transmitter applications, the DBM is most readily recognized for its role as a balanced modulator. If an RF signal is applied to the RF port with no signal applied to the IF port, then no signal appears at the LO port. The DBM's native state is to suppress an RF carrier in this application. If a modulating ac signal is then applied to the IF port, the DBM is periodically unbalanced, allowing alternately-phased envelopes of RF to pass through the device. The result of this is a double sideband suppressed carrier (DSBSC) signal (**Figure 6**).

Now, while DSBSC is not commonly used in amateur radio (although my ancient Central Electronics 100V transmitter can produce a top-notch DSBSC signal, along with several other obsolete modes), there are actually some very useful applications of the mode, primarily for scientific applications. We'll discuss some of these in detail later.

For most modern-day radio amateurs, DSBSC is considered an interim stage in producing what we really want, which is a single sideband suppressed carrier (SSBSC) signal. But before going into SSB too far, let's back up and work with the DBM itself some more.

Full or Partial Carrier AM

We described earlier how the DBM can be used as an attenuator. Or, looking at it a bit differently, we can consider it a classic amplitude modulation block. By applying a dc bias to the IF port, we can allow a portion of an RF signal to pass through the device, in other words a carrier. Now, if we superimpose a modulating signal on that dc bias (**Figure 7**), we now have a full-carrier AM signal (**Figure 8**). The nice feature of creating AM this way is that it allows us to very simply adjust the carrier level by adjusting the dc bias. Or if we want to be even more elaborate, we can even incorporate the dc bias inside a feedback loop, controlled by the average modulation, and produce what is known as controlled-carrier AM, a power-saving method for high-power AM operation. While this method is not exactly new — it has been used in high-power shortwave stations for decades — it's a lot easier to implement with our versatile DBM block. All the modulation takes place at low power levels. The price to pay for this convenience is that any amplification after the modulation scheme must be done with a linear RF amplifier at some sacrifice of efficiency compared to high-level AM modulation methods.

Compound Modulation

What if we were to take our DBM-based balanced modulator, and in turn modulate that with a frequency modulated signal? Can you actually do such a thing? You certainly can. In this case it would be a frequency modulated DSBSC signal. The use of modulated subcarriers to, in turn, modulate other carriers is a common commercial radio practice. There's no theoretical limit to the number of subcarrier layers you can assemble together, but there are some practical limits — and probably regulatory ones as well.

Phase Detection

One of the interesting uses of the DBM is as a phase detector, or more accurately, a phase comparator. If you apply an RF signal of the same frequency to the RF port and the LO port, the IF port will give you a dc voltage proportional to the phase difference between the signals. The dc voltage will be 0 or 'centered' when there is a phase difference of 90°, with a fairly linear change of dc voltage as the relative phasing of the two signals varies from 0° to 180°. In a typical configuration, the RF phase of an incoming radio signal will be compared to a stable local oscillator.

When used as a phase detector, the DBM needs to be operated in a fully saturated mode, or the signal needs to be limited beforehand, if you want to determine *only* the phase difference. The DBM is sensitive to *both* the phase angle and the RF amplitude, which will give you ambiguous results on amplitude varying signals. But what if you want to know the phase *and* the amplitude of in incoming signal?

In this case, you will want to use *two* phase detectors driven in quadrature at their LO ports. This is the configuration of the lock-in amplifier, the venerable stalwart of extremely weak signal detection. In this device each of the DBMs is both phase and amplitude sensitive, but by using a known I/Q reference, you retain the phase information.

Synchronous Detection

There are two problems with the conventional method of using a passive diode detector for AM reception. Number one is that diodes have a threshold below which they fail to detect, or do so with a lot of distortion. Number two is that, at least for the single-ended diode detector, there is a particularly obnoxious form of distortion known as diagonal clipping, or failure-to-follow distortion. This defect has been well known for many decades, and is primarily addressed by careful selection of time constants and IF bandwidth, all being compromises to some degree.

As hams discovered quite a while ago, using a BFO — normally reserved for CW and SSB detection — also enhanced AM performance of receivers, especially in the area of sensitivity. However, the BFO had to be placed *precisely* on the incoming AM carrier frequency — a difficult job with the hardware of the time. The DBM's role as a product detector solves this problem in style. The DBM allows a simple and effective means of employing synchronous detection, the ideal way to demodulate AM, without the need for exotic phase locked loops or the like (see **Figure 9**). First the AM IF is amplified, run through a narrow filter centered on the carrier (relative to the IF), limited, and then applied to the LO port of a DBM. The unlimited, unfiltered, 'raw' IF signal is applied to the RF port. The IF port is now the detected audio output. This audio is exceptionally clean and stable, and fairly immune to selective fading as well.

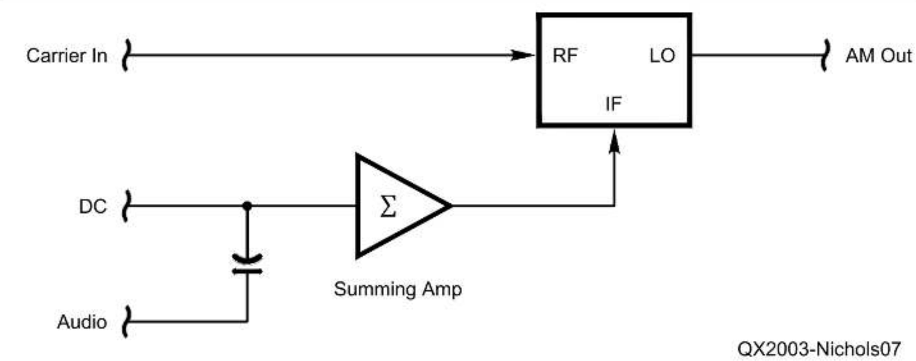


Figure 7 — A balanced mixer can be used to generate every form of AM, from full carrier to fully suppressed carrier DSB, by the simple addition of a dc offset (bias) applied to the modulation input of a DBM.

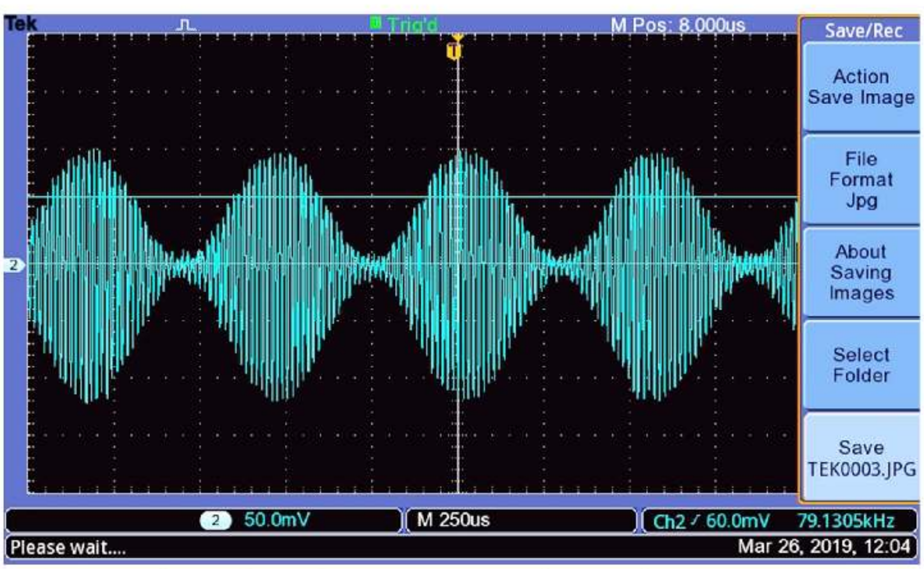


Figure 8 — A full-carrier AM signal created by applying a dc bias voltage to the DBM as well as the modulating waveform. Modulation depth is nearly 100% in this example.

Some Non-Radio Applications

I first heard about ring modulators in reference to music synthesizers, particularly the famous Moog Synthesizer in the late 1960s, which paved the way for just about every kind of electronic music to follow. In addition to imitating and simulating a vast variety of existing musical instruments, the Moog was capable of generating sounds that had never before been experienced in nature. Music tends to be harmonically related; electronic instruments up to that point worked by filtering out complex waveforms with band-pass filters and such where Fourier analysis still reigned supreme. Pipe organs, for example, could be closely — or fairly closely) — duplicated by the judicious filtering of harmonically generated electrical signals. The advent of the ring modulator, however, created sounds based on arithmetic addition, rather than frequency multiplication, which created some extremely unusual and sometimes disturbing audio experiences. Musical notes could be ring-modulated with other musical notes, or even noise, to create entirely new sounds that didn't exist on Earth (and probably not on Mars). You can get an idea of what ring modulation sounds like by listening to a shortwave music station with a slightly off tuned SSB receiver. It can be anything but musical, or what we might consider musical. Used sparingly and carefully, however, ring-modulated music can still sound somewhat pleasant, if unusual. In any case, the ring modulator is one of the main ingredients that set modern music and sound production on its present course, for better or for worse.

The double balanced mixer also has application in optics, though in that particular application the hardware is not immediately recognizable as such. However the functional equivalent is often used for up-converting and down-converting optical frequencies. For instance, high powered lasers for nuclear research are replete with numerous optical frequency converters, using non-linear optics — diode-like devices — to do the work.

Rolling Your Own

While the modular DBM is readily available from a number of sources, such as Mini-Circuits, it can be quite informative to build your own DBM for experimentation. Like so many of my projects that begin innocently enough with the acquisition of an intriguing surplus component, I stumbled across a few pulse transformers, which worked beautifully as the basis for a couple of broadband homebrew DBMs (Figure 3). I also show the DBMs as connected for generating the oscilloscope displays in **Figure 10**. As you can see, miniaturization was not a consideration. I simply wanted a circuit that I could poke around on and see how these devices *really* worked. These are really great for generating and demonstrating a number of modulated waveforms. Eventually, these will end up in a lock-in amplifier that I'm putting together for 2200 meters.

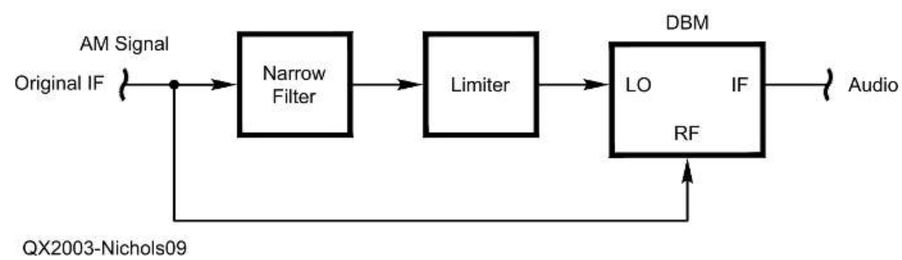


Figure 9 — A "Poor Man's Synchronous AM Detector" can be used to reduce distortion caused by fading. More elaborate synchronous detectors using phase locked loops are now common in high quality receivers.

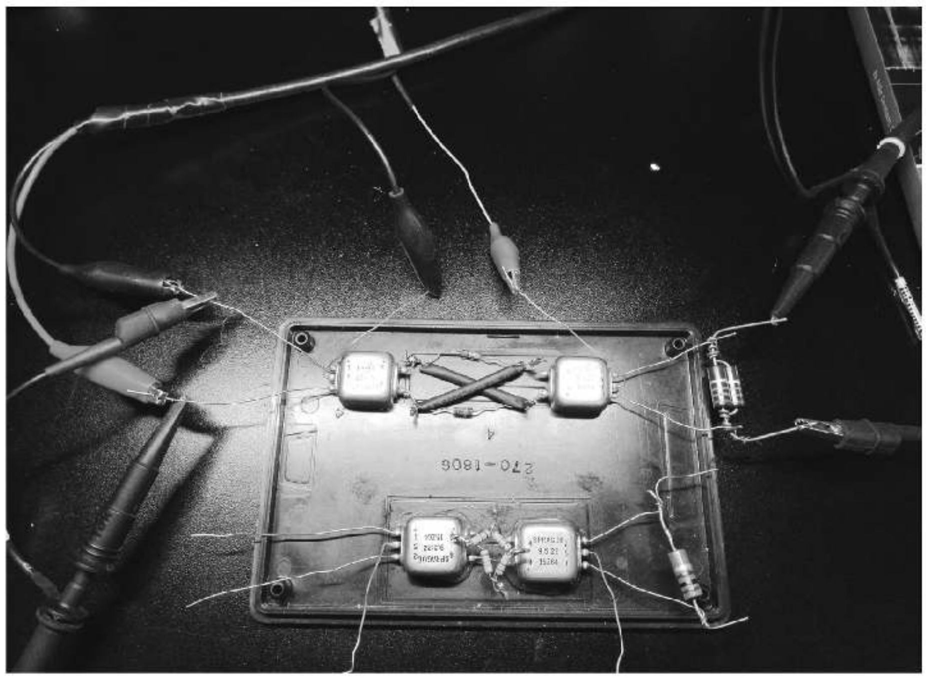


Figure 10 — DBMs work best when operated into a specific value non-reactive load; the prototype DBMs here are no exception. While commercial DBMs are usually designed for 50 Ω load impedances, this is by no means the only possibility. The ready availability of a wide variety of ferrite cores allows the creation of input and output transformers of any impedance value.

Etc.

There are probably a few — or maybe many — other applications for the DBM that I haven't thought of. Hopefully this has triggered a few ideas for different ways to use this many-faceted building block. I highly recommend looking at the DBM application notes on the Mini-Circuits web page (**https://www.minicircuits.com/applications/application_notes.html** for practical details on using prefabricated DBMs.

There are a few other manufacturers of very similar items, but I believe Mini-Circuits has the widest and most economical selections. I always try to pay a visit to their booth at Hamvention.

Eric P. Nichols, KL7AJ, is a two-time recipient of the William Orr, W6SAI, Technical Writing Award. He has written numerous articles for QST and QEX magazine as well as four ARRL books, the latest being "Receiving Antennas for the Radio Amateur." Eric's latest focus is on encouraging experimentation on our two new bands on 2200 meters and 630 meters, heroically attempting to make up for lost time in catching up with our LF brethren "across the pond.

Ralph J. Crumrine, NØKC

1621 E. Sunvale Drive, Olathe, KS 66062-2108; **n0kc@arrl.net**

A 3 kW, 65 V Light Weight PWM Power Supply

This power supply was designed to support state-of-the-art RF LDMOS FET amplifiers.

When I designed the DIY-3300 high voltage supply for legal power limit vacuum tube amplifiers (see the *QST* article **[1]**) it seemed that my lack of interest in high power solid state amplifiers would keep me from ever considering power supplies for such equipment. My reasoning was: 1) I haven't much interest in transistor amplifiers, and 2) probably power supplies for that purpose would be plentiful if I should ever need one. I was wrong on both counts.

The recent announcement by NXP of a 65 V LDMOS FET device caught my eye. To experiment with that device in a linear amplifier was going to require some considerable power at +65 V. There is a scattering of 50 V supplies available that would require modification, if possible, to suit my needs. These supplies are dated in their selection of components. I thought it would be interesting to do some design work with some state-of-the-art switching MOSFETS and come out with a ready result of 65 V output with some features particularly useful to the radio amateur operator/experimenter.

Currently available stand-alone supplies of this sort are all design centered at 50 V for typical industrial and older RF power transistor requirements. Also these supplies could use some updating with new components. In one offering for instance, two supplies of old design were paralleled to get the output power needed at 50 V dc. Size and weight reduction was a big consideration *à la* the DIY-3300. The resultant new design is exactly the same mechanical size as the DIY-3300, using the same sheet metal for the cabinet, and same dimensions for the PC board, see **Figure 1**. It weighs about 9 lbs, a bit heavier than the HV supply because it has a filter inductor as well as a transformer.

Figure 1 — The power supply weighs about 9 lbs.

The latest of the new MOSFET switching devices are cast from GaN (gallium nitride) semiconductor material. The device in mind for use here was limited in drain voltage and, therefore, was amenable only to bridge circuit configurations **[2, 3]** when operated from a 240 V ac line. To cut down on the size, weight, and cost of power electrolytic capacitors needed for the half-bridge circuit, I selected the full bridge circuit.

The GaN devices have extreme power gain and bandwidth, and that proved to be a real challenge finding the refinements in the bridge circuit to keep them from blowing up. Even though they are driven as ON-OFF switches, they are capable of breaking into oscillation during the switching transient, with severe oscillations at UHF. After several generations and a number of explosions, I decided to go back to the tried and true 1,200 V SiC MOSFETs of the HV power supply. This has the advantage of reducing the number of MOSFETs from four to two as required for the push-pull circuit topology. The resultant circuitry, then, is a push-pull transformer isolated, buck mode [2], pulse width modulated switching power supply.

The power supply design described here brings the weight of a 3 kW supply down to nine pounds. The same ferrite core pieces used in the DIY-3300 are used throughout this design for the power transformer and filter inductor. The design operates at the same 60,000 Hz switching speed. This is not so fast switching as to reach into the exotic realm of parts and cost. It is sufficiently fast to get a handsome reduction in size and weight. See the DIY3300 article [1] for a narrative with pictures on what this switching speed can do for size and weight reduction.

Simplifying the construction can help in some small way to reduce weight. This project is built up almost entirely on a single printed circuit board. Anyone can build this power supply if they can stuff a printed circuit board. See **Figure 2** for a view of the completed PC board installed on the chassis.

Table 1.
Power and voltage output measurements, with Vac in = 241 V rms 60 Hz

Idc, A	*Vdc, V*	*P, watts*	**Vac rms, mV*	*(Vac/Vdc), %*	*(Vac/Vdc), dB*
1.6	66.4	106	49	0.07	-62.6
12.0	65.7	788	90	0.14	-57.3
23.7	65.6	1,555	121	0.18	-54.7
35.4	65.5	2,319	154	0.24	-52.6
47.2	65.4	3,087	179	0.27	-51.3

**Measured with Fluke 89 IV DMM, bandwidth = 500 kHz.*

Figure 2 — Top view of the completed PC board installed on the chassis.

Features and Advantages of this Power Supply

1) *High speed, high voltage switching using the latest SiC MOSFETs*. Ultra high speed recovery, low voltage drop dual diode arrays are used for rugged high output current. These parts allow the high switching speed of 60 kHz. At such switching speeds silicon IGBT transistors and high current UF rated silicon diodes these days are really considered marginal in their performance.

2) *Voltage regulation*. Voltage regulation is provided for by pulse width modulating (PWM) control in a stabilized feedback control loop. See **Table 1** for the quality of regulation achieved under load.

3) *Other output voltages up to +65 V are possible*. See the section describing the regulation and feedback control.

4) *Outstanding size and weight reduction*. For a 3 kW power output, the size is 5 1/2" H by 7-1/2" W by 10" D and weighs only 9 lb.

5) *Output isolation*. Output is fully isolated from the ac line with the negative side of the 65 V supply connected to chassis to provide a positive voltage supply to an N-channel MOSFET amplifier load.

6) *Selectable mode of over-current protection*. An internal switch selects fast shut-down or current limited continued operation over-current protection. A front panel LED announces an over-current condition. In the shut-down mode the LED is lit and the reset button must be pushed to start a re-try. In the continuous mode the LED lights at about 75% of over-current limiting. In full limiting, the supply voltage is reduced to hold the current to less than the limiting maximum.

7) *Overheating shutdown protection*. Hot spot monitoring by a heat sink mounted thermistor causes shut down if necessary.

8) *Extremely low output filter energy retention*. Should a short occur and the supply shut down, the output filter circuit will dump less than 1/1000 of the energy stored in an equivalent 60 Hz design into the short.

9) *EMI considerations*. A complete metal enclosure is used with RF filtering on all connections in and out of the enclosure to minimize interference from the supply. As a further feature of EMI control, the supply may be operated in the passive mode. It can be keyed on by the TX key line (see item 13 below) from the transceiver, hence there is no noise output between transmissions, that is, during the receive mode.

10) *Auxiliary supply*. Regulated +24 V auxiliary supply and a programmable supply up to a maximum of +14 V at up to a total of 2 A are included to completely power almost any linear amplifier. If not needed, the programmable supply can be omitted for a minor cost savings. Also, the transformer can be reduced from 56 VA to 35 VA for an additional cost savings.

11) *Remote operation*. Control of the basic functions of the power supply can be done remotely. Turning on the power supply can be done with either 24 V ac or dc signal at 10 mA, either two wire isolated or one wire positive with respect to chassis ground.

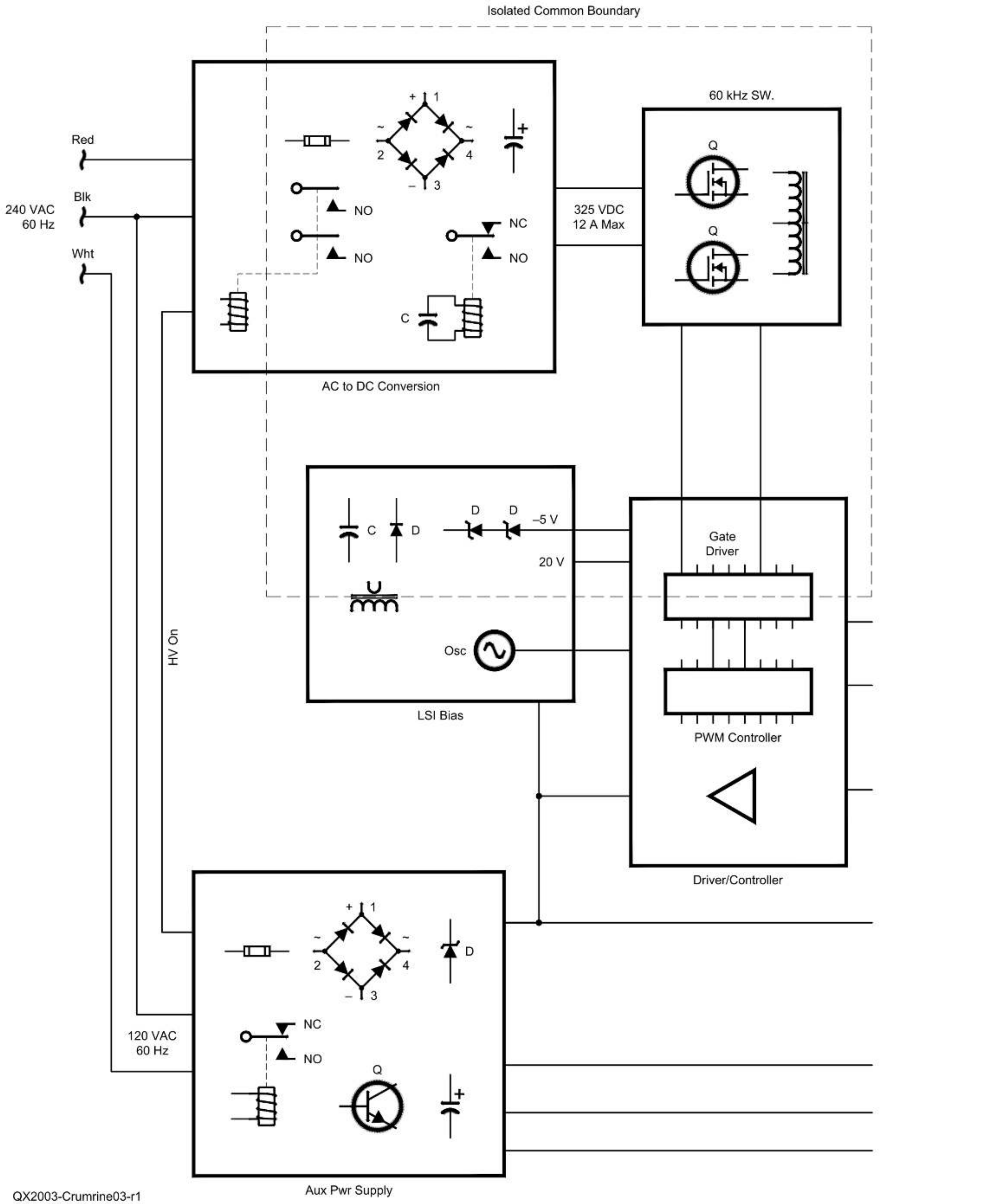
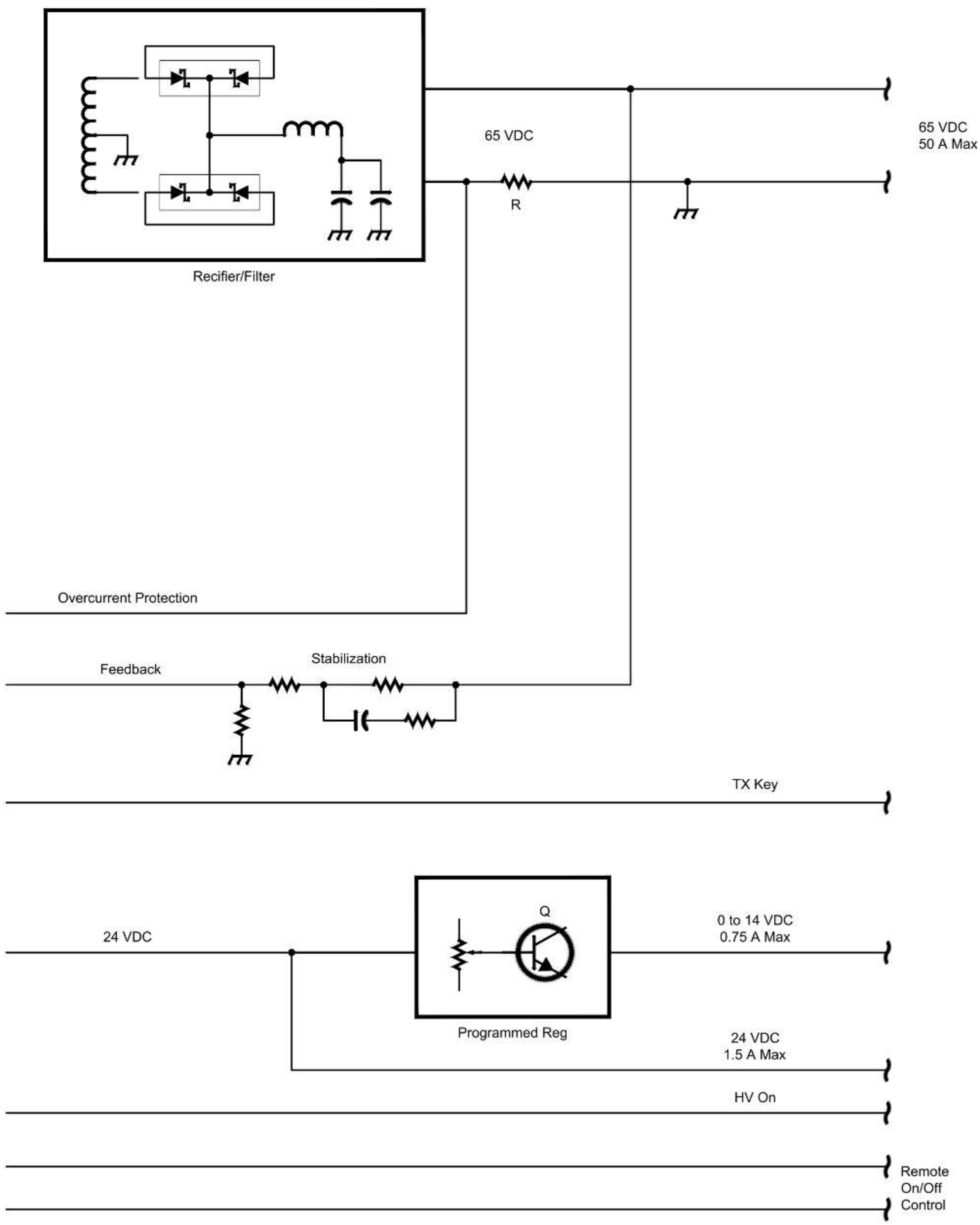


Figure 3 — Simplified block diagram of the power supply.


65 VDC
R
Recifier/Filter
65 VDC
50 A Max
Overcurrent Protection
Feedback
Stabilization
TX Key
Q
24 VDC
0 to 14 VDC
0.75 A Max
Programmed Reg
24 VDC
1.5 A Max
HV On
Remote
On/Off
Control

Turning on the +65 V output remotely requires a switch connection to chassis ground conducting 30 mA.

12) *Dual soft start features*. In the ac primary power circuit a soft start circuit that uses feedback involving the primary supply bleeder resistor limits the inrush current from the ac mains. A second soft start circuit provided by the LSI controller chip provides pulse width slewing starting with zero pulse width. This circuit takes effect whenever the +65 V output is energized, be it by the LV power switch, the over current reset switch, or the transmitter TX key line.

13) *A soft start option*. A soft start via the MIC key line is an option by removing a diode. The rise time is such that this is appropriate to AM, SSB or FM voice operation. It is not fast enough to be used with any sort of break-in operation such as CW break-in regardless of break-in delay. By using the TX key line the +65 V supply is inoperative during receive conditions, free of radio noise in the receive mode.

14) *Cool, quiet operation*. Cooling is completely within the power supply assembly shown here. No further addition of external heat sinks and fans is necessary. A small 92 mm square 24 V dc box fan operates at half rated voltage from a cold start. From that cold start the fan speed is controlled by a servo loop to apply just enough air to control heat rise, keeping the fan noise at the lowest level possible. A plug-in provision is made for a second fan to be added in tandem at the rear of the unit for extra heavy duty use or use in high ambient temperatures, and/or continuous FM or teletype broadcast.

Regulation and Feedback Control

The regulated +65 V output is achieved through the use of the TL494 PWM controller IC. The +65 V output is fed back to the TL494 through a divider and stability compensation network. It is possible to adjust the divider to ratios that will produce lower output voltages, but not higher.

Magnetic Components Design

The transformer is a double stacked, four-piece E core set similar to the DIY3300. This design would probably be overkill for a commercial application. But it has proven to give the quality of coupling comparable to a fully enclosed ferrite Pot (P) core, with a coupling factor of better than $k = 0.996$, resulting in the lowest leakage reactance possible **[4]**. The filter inductor uses two of the same E cores in a gapped design to handle the dc current involved. The bodies of the transformer and filter inductor are wrapped by an external shorted turn of copper foil greatly reducing radiation from these components.

The Simplified Block Diagram

Circuit operation can be understood from a look at the block diagram in **Figure 3**. The full schematic along with additional information is available from the **www.arrl.org/QEXfiles** web page. We have:

AC to DC Conversion

The 240 V ac line is converted to an isolated 325 V dc supply through the use of a diode bridge rectifier array and capacitor filter. A DPST relay at the entry point of the 240 V ac line is used to connect this high voltage to the rest of the circuit. An SPST relay is used to switch out the soft start current limiting resistor. The two relays form the primary soft start circuit. The first relay is closed by the LV ON control. The second relay, for the moment is open, routing the inrush current through a limiting resistor. With the charging of the primary supply capacitor nearly complete, the second relay is closed to cut out the limit resistor. This is done by the current in the bleeder resistor and delayed by a timing capacitor across the coil of the second relay.

60 kHz Switching

The 325 V dc is applied to a push-pull arrangement of transformer windings and two switching transistors. The switching transistor gates are driven by an LSI gate driver that can supply the turn-on pulse and turn-off reverse biasing for best control of the switching transistors. Half of the center-tapped secondary winding produces a peak output of ±85 V.

Rectifier/Filter

The secondary winding drives two diode arrays as a center tapped full wave rectifier. The rectifier output is followed by the "buck" inductor to bring the output down to a constant +65 V filtered by a pair of electrolytic capacitors. A bleeder resistor is in this section, but not shown.

Over-Current Protection

Output current is sensed across a 1.4 mΩ resistor printed circuit path that connects the negative (-) output terminal back to the center tap of the transformer output winding. This produces a negative going voltage as a function of output current. This voltage drives a comparator and op-amp providing either latched shut down or continuous current limit. A front panel LED provides a warning of an over-current condition.

Driver / Controller

The two pulse trains of proper time period and interleaving are provided by the controller IC. The pulsed outputs from the controller are then applied to the LSI gate driver that provides the necessary isolation between the controller IC and the 325 V dc switching circuit, avoiding the use of driver transformers. These pulse trains provide for a complete OFF condition between pulse trains. Further protection against any possible overlap of the two pulse trains is done by logic in the LSI gate driver chip. The pulse widths vary according to the load to regulate the output voltage at +65 V.

LSI Isolated Supply

The secondary side of the gate driver LSI requires -5 and +20 V isolated supplies. These voltages are supplied by a micro power switcher and isolating fly-back transformer circuit. The drive to the switch is taken from the 60 kHz oscillator in the PWM controller LSI.

Internal and Auxiliary Power Supply

A 120/240 V ac power transformer with bridge rectifier, and capacitance filter followed by a +24 V regulator IC provides that voltage as an output and drives another regulator IC set to any voltage up to +14 V dc at its output. These voltages are provided for external use in control of the amplifier connected to the power supply. The programmed voltage is not used in the DIY6550 proper.

Special Considerations

When the 240 V ac power service is bridge rectified, as it is here, the rectifier output and associated circuitry of the power switching transistors must be totally isolated from all other components that operate from a chassis-connected power source. This isolated circuitry is shown in the block diagram, Figure 3, as all components inside the dashed line boundary. Any test equipment connected to this circuitry must be fully insulated from power ground. That is, it should either be battery operated or ac-line operated through an isolation power transformer.

Construction

This project can be completely done from the package of drawings, schematic, and bill of material that are available. The design is intentionally done so that it is nearly completely assembled on a single printed circuit board, see Figure 3. When the board is stuffed, you are very nearly finished with assembly.

The idea of building a transformer may be a put-off to some Do-it-Your-Selfers. Actually, it is kind of fun to build your own. The transformer core is made up of two pairs (4 pieces) of E-shaped ferrite material. This small transformer uses ready-made bobbins that match the cores, and the turns count is so low that hand winding the bobbin becomes not just practical but easy. The single winding of the filter inductor is even simpler.

For a discussion of methods used in completing the enclosure, sheet metal fabrication, painting and lettering, I would direct the reader to my article on the Centennial Amplifier **[5]**.

Those interested in this project should contact me. With enough interest I would serve as a source for the special parts such as the transformer, inductor, and printed circuit board, etc. All other materials can be found through the usual electronic parts suppliers and eBay. Direct any questions to me at **n0kc@arrl.net**.

Ralph J. Crumrine, NØKC, was first licensed as a Novice in 1953 as WN3WFZ. He upgraded to General class and finally in 1978 to Amateur Extra class. His enlistment in the USAF put him to work in radio and navigation equipment repair. After military service he attended The Pennsylvania State University where he earned a BSEE degree, graduating with honors. A career followed in the design and development of avionics equipment, beginning work with King Radio Corporation and finally retiring from Honeywell Avionics Division. Ralph is a member of ARRL and has been an active ham in retirement, earning the WAS and DXCC awards in 2002. Antenna and equipment design, and design practices and procedures have been of particular interest with several articles published on these subjects with the ARRL.

Notes

[1] R. J. Crumrine, NØKC, "A Small, Lightweight High-Voltage Switch-Mode Power Supply", *QST*, Jan., 2017, pp. 30-34.

[2] A. I. Pressman, *Switching Power Supply Design*, McGraw-Hill, 3rd edition, 2009.

[3] B. Andreycak, et al., *Unitrode Switching Regulated Power Supply Design Seminar Manual*, SEM700, 1990, Unitrode Semiconductor Products Division.

[4] N. R. Grossner, *Transformers for Electronic Circuits*, McGraw-Hill, 1967

[5] R. J. Crumrine, NØKC, "A 1500 W Centennial Amplifier for the 80 - 6 Meter Bands," *QST*, Dec., 2014, pp. 30-35.

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HM-133 Microphone DTMF Adapter Project

Add DTMF capability to popular transceiver microphones.

This project uses a microcontroller and some support circuitry to allow an Icom HM-133 microphone to be used with any transceiver with the use of an appropriate adapter cable. The circuit accepts the keyed data stream and interprets the keys. If a key is in the DTMF pad-space, and the PTT is active, the microcontroller generates the appropriate DTMF tone. The PTT, UP, and DOWN buttons are also interpreted and provided as open-drain outputs to the radio connector.

Jumper and component options provide a great deal of flexibility with regard to the radio interface. In addition, several 'key-macros' are provided to allow a given number of pulses to be issued on the up or down button outputs.

The adapter is contained on a small circuit board (**Figure 1**) that is placed in-line with the HM-133 to radio connection. The HM-133 and associated cable require no modifications. RJ-45 jacks provide the HM-133 and radio connections. An appropriate cable is fashioned for the desired radio that consists of an RJ-45 plug and the appropriate radio connector (typically, an 8-pin DIN style). The adapter and HM-133 use 8 V for power from the target radio. The system is compatible with virtually any transceiver that can accept a DTMF microphone.

The DTMF Adapter

I have an Icom IC-7000 that uses an HM-151 microphone. As capable as the radio is, it cannot generate DTMF signaling with the HM-151. This seems like a significant oversight for a radio that can operate on FM repeater frequencies in the VHF/UHF spectrum.

Recently an acquaintance of mine who shares my affinity for the Icom IC-901 had a problem locating a DTMF microphone for his radio. That got me thinking, if I needed a DTMF microphone, what options might I have? I have several Icom HM-133 and HM-151 microphones, so I wanted to see if they could be adapted to create DTMF signals.

A few years ago, I was interested in decoding my HM-151 key-data output. I was aided by John Wren, K4JCW, regarding the HM-133 microphone. He had posted helpful material on his Wiki page [**1**]. The HM-151 is very similar to the HM-133, allowing me to write a decoder application for an ARM-based radio controller I have been working on for some time. Thus the code for this was essentially done. Direct Digital Synthesis (DDS) tone generation was also an easy task. I chose a small 8051 MCU, and designed the necessary circuits to capture the microphone data, merge the DTMF signals with the microphone audio,

Figure 1 — The HM-133 DTMF adapter assembly.

and squeeze it all into as small an area as practical. The resulting DTMF adapter is seen in Figure 1. The schematic circuit is on the **www.arrl.org/QEXfiles** web page. The DTMF adapter operates as a dongle connected between the microphone and the radio (**Figure 2**).

A comparator peripheral built into the MCU captures the serial data from the microphone. This is then fed to a timer pin that captures the bit edges for decoding in software. A pair of timers generate a PWM digital-to-analog converter and drive the DDS interrupt. FET transistors create PTT, UP, DOWN, and FUNCTION signals. An analog switch steers the tone and mutes the microphone while DTMF signaling is taking place. For the HM-151, another analog switch interrupts the data path to allow the DTMF keys to be pressed without sending potentially confusing data to the down-stream radio. A voltage regulator, EMI ferrite beads, and some components to support in-circuit programming of the MCU round out the circuits.

The Software and PCB

The software operates in a single loop that polls for serial data from the microphone and parses it into actions — either tones or GPIO pulses. The software supports UP, DOWN, and 4 FUNCTION buttons as well as the DTMF keys. The FUNCTION buttons map to open-drain outputs that each have a series resistor of a particular value that can be installed. The available options allow the interface to mimic the operation of the Icom HM-14, Kenwood MC-44DM, and Yaesu MH-36 microphones. Strapping resistors allow the microphone/radio emulation to be configured, allowing a single software load to support the available options. Currently 6 of 8 possible options are assigned, with the remainder available for future options.

I used **PCBgogo.com** for inexpensive prototype PCBs. For about US$25, I got 5 PCBs delivered in just 5 days. The PCB features some rather small components. This could be a serious impediment for some. Thus, I expected that a partial kit might be made available — assuming there is enough interest in this project — that limits the number of components that one needs to solder, and also limits these components to packages that are at least manageable for those who are able to deal with components that are 0603-size or larger. I have created a web page **[2]** to act as a clearing house for more detailed assembly

Figure 2 — The DTMF adapter assembly operates as a dongle. The final assembly would be covered by a 3D printed enclosure or by a large-diameter heat shrink sleeve.

and operating instructions and software programming. The schematic and additional files are available on the **www.arrl.org/QEXfiles** web page. The latest availability status on PCB or partial-kit options can also be found there.

A short RJ-45 cable connects to the target radio. The connector for the radio is wired to the appropriate radio RJ-45 connector pin. For feed-through applications, such as the IC-7000, a short 1:1 RJ-45 cable may be used. Be mindful of the signal pairing if CAT-5 cable is used, however for short runs (less than 6 inch), it shouldn't make a lot of difference. In general, keep the interface cable as short as practical.

The HM-151 interface operates as one might expect. Simply push the PTT and when you press any of the DTMF keys the appropriate DTMF signal is presented to the microphone audio line, and the microphone is muted for about 1.5 s. When configured for the IC-7000, the HM-151 F-2 button is used to toggle the system in/out of DTMF mode. This suspends the data-path to the radio so that the DTMF digit data does not interfere with the IC-7000 operation.

When using the HM-133, you must place the microphone into DTMF mode using the DTMF S button. PTT is *not* pressed, simply press the DTMF digits and the PTT is activated automatically. This is dictated by how the HM-133 operates and I've not found a satisfactory means around that operating method, other than modifying the HM-133 to use a hard-wired PTT as in the HM-151. This is easy to do, but the HM-133 will likely no longer function on a radio designed for it.

The software features an UP/DOWN macro system for generic radio interfaces. If you enter one or two numeric digits (0-9) with no PTT (or DTMF-S OFF), and then press the UP or DOWN button, the interface will pulse the respective output up to 99 times. Also, holding the UP or DOWN button for more than 2 s will cause the interface to pulse the respective output until the button is released. An optional LED output is provided that is routed to HM-133/151 pin 2 UP/DOWN, that is not connected on these microphones. Careful placement of a suitable LED on the HM-133/151 PCB allows this LED to offer feedback for the macro operations. The LED pulses for key presses or is OFF while the microphone is muted.

Joseph M. Haas, KEØFF, holds an Amateur Extra class license. He started tinkering with electronics at the age of 6 and was first licensed in 1978 as KAØGPZ. Joseph earned a BSEE from the University of Missouri, Rolla (now the Missouri school of Science and Technology). He currently works in the SAT-COM industry. He has worked in avionics design, oil-field instrumentation design, and semiconductor process equipment design. Most of his professional and hobby designs are mixed-signal in nature. They involve a microcontroller performing as a control system and/or data converter for signals from dc up to a few megahertz. He has also designed RF circuits in the 2.5 GHz range and has experimented at 10 GHz. Joseph maintains a projects page at **www.rollanet.org/~joeh/projects/**.

Notes

[1] J. C. Wren, K4JCW, **tinymicros.com/wiki/Icom_HM-98/HM-133_Internals**. (Online: 8/19/19).

[2] J. Haas, KEØFF, **www.rollanet.org/~joeh/projects/hm133/**. (Online: 8/27/2019).

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Collection of Broadband HF Antenna Designs, Part - 1

A wide range of new broadband wire antennas results from an application of optimization software to antenna numerical analysis software.

[This article is presented in three parts. This Part – 1 contains Sections 1 through 7 and includes only Figures of antenna details, while a later Part – 2 contains Section 8 through 10 with antenna detail Figures. Part – 3 **QEXfiles** available online contains all of the Figures, including those that show antenna performance details. The Figures of Parts – 1, 2, and 3 are sequentially numbered. — *Ed.*]

1 - Introduction

The number of broadband antenna designs for amateur HF bands is lower than the number of narrowband ones. Log-Periodic Dipole Array, Tilted Terminated Folded Dipole, Conical Monopole or Discone come to mind but not many more. And yet, when your available space makes it impractical to build a number of monoband antennas or a tedious trimming of a multiband resonant antenna is something you would like to avoid, a single antenna covering 14 - 29.7 MHz frequency range in a continuous way is very desirable. Thanks to antenna simulation programs available today, it is possible to design such antennas. The antenna optimizers that can do the optimization not only for a single frequency but for a number of frequencies at the same time are especially convenient for the purpose. The *AutoEZ*, an add-on to *EZNEC* program, has such an optimizer. I used it intensively to finalize many antenna designs described here and in Part – 2 of this article.

2 - Initial Assumptions

Before moving on to antenna descriptions, we must answer two questions:

1) what must be the operational frequency range of a broadband antenna?

2) how large a standing wave ratio at the antenna feed point (SWR_{ANT}) is acceptable in the operational frequency range?

The answer to the first question is: the antenna must cover 14.0 MHz to 29.7 MHz range. Keeping F_{MAX} to F_{MIN} ratio only slightly bigger than 2 (but not greater) prevents getting multi-lobe radiation patterns that are usually not desired.

The answer to the second question is: SWR_{ANT} smaller than 3:1 throughout the frequency range will be considered good, SWR_{ANT} smaller than 4:1 will be considered acceptable.

Why such numbers? It is assumed that the coaxial feed line has matched line loss 2 to 3 dB over the 14.0 MHz - 29.7 MHz frequency range. Such values are typical for about 30 m or 100 ft of RG-58 coax. For longer coax runs, you would most likely use a lower loss cable like RG-8 or RG-213 to get no more than 2 to 3 dB of loss.

Now, if the SWR_{ANT} at the antenna feed point is 3:1, then the additional loss in the feed line due to *SWR* will be about 0.8 dB for 14 MHz and 1.0 dB for 29.7 MHz.

For SWR_{ANT} = 4:1, the additional loss will be approximately 1.2 and 1.5 dB for 14 and 29.7 MHz frequencies respectively.

A 'dBdm' [that is, a modified dBd, where the usual 0 dBd = 2.15 dBi — *Ed.*] unit will be used in this paper to compare the broadband antenna with its feed line (and extra losses due to *SWR*) versus a half-wave dipole with exactly the same feed line but without extra loss due to *SWR*. It will be assumed that the reference dipole has SWR_{ANT} 1:1. In this way, we will never be overly optimistic about the performance of any broadband antenna design presented here. In other words, in order to achieve 0 dBdm, the broadband antenna under analysis will need to have gain *G*,

$$\begin{aligned} G &= 2.15 \text{ dBi} + x \\ &= 0 \text{ dBd} + x \\ &= 0 \text{ dBdm} \end{aligned}$$

where x is an extra loss in the feed line due to *SWR*. For example, if the SWR_{ANT} = 3:1 and the frequency is 29.7 MHz, the antenna will need to have gain: 2.15 + 1 = 3.15 dBi to achieve 0 dBdm.

In all the above estimations, the SWR_{ANT} at the antenna feed point was taken into account. Please keep in mind that a lossy feed line causes the SWR_{TRX} measured at the transceiver end of the coax lower than SWR_{ANT}. With a coax cable as above, you may expect something like SWR_{TRX} < 2:1 for SWR_{ANT} <3:1 and SWR_{TRX} < 2.5:1 for SWR_{ANT} < 4:1. Any transceiver equipped with automatic antenna tuner will be able to match its output to the broadband antenna with its feed line and there will be no extra signal loss in the transmitter output stage.

Multiband resonant antennas, especially the compact ones, quite often have high SWR_{ANT} at the edges of some amateur bands — sometimes as high as 5:1. Therefore, setting SWR_{ANT} = 4:1 as a limit for a broadband antenna is quite reasonable.

Most of the antennas described in this article were simulated in free space because such results are more convenient if you want to compare performance of different antennas. The monopole antennas that require RF ground for proper operation were simulated with a system of radials over a high accuracy average ground model.

Ideal lossless wires of diameter 2 mm were used in all models described in this paper. Before building a real antenna, a given antenna model should be edited for the intended kind of material (Cu, Al, Fe) and element diameters should be set to the expected values. It would be also a good idea to simulate the antenna model over the ground at the intended installation height rather than in free space because the presence of ground will influence SWR_{ANT} a bit. Small adjustments of antenna dimensions may be required to achieve the best possible performance.

3 - Terminated Folded Dipole

Let's start with something well-known. You probably know this antenna under the name Tilted Terminated Folded Dipole, TTFD or T2FD. As the tilt will not be present in a rotatable antenna, I omitted this descriptor in the antenna name. After using an antenna simulator for a number of feed impedances, I obtained the best results for the source resistance $Z_0 = 600\ \Omega$. The antenna is shown in **Figure 1** and its dimensions are as follows: length is10.28 m, distance between the antenna wires is 0.2 m, termination resistance is 900 Ω.

The plots in **Figure 2** (see the **QEXfiles** web page) show SWR_{ANT}, gain and radiation patterns of the TFD. The TFD has a low SWR_{ANT} throughout the entire frequency range with a maximum value of 2.5:1 at 17.2 MHz. The shape of its radiation patterns is bidirectional and leaves nothing to be desired. The only problem is TFD gain. Because a significant amount of power is dissipated in the termination resistor, the power converted into electromagnetic wave is smaller than that of a dipole. At 14 MHz, the TFD gain is -3.7 dBi. And because its SWR_{ANT} is about 2:1 at that frequency, the relative gain of TFD is -6.3 dBdm.

The TFD gain improves with frequency and rises to -1.1 dBi at 29.7. But it is still -3.3 dBdm when calculated with the feed line. In other words, depending on frequency, the TFD with a feed line produces 3.3 through 6.3 dB weaker signal than a dipole with the same feed line.

It is worth knowing that if 100 W is applied to the TFD feed point, 74 W is lost in the resistor at 14 MHz, and 70 W is lost at 29.7 MHz. I think everybody will agree that this is too much and it really makes sense to look for better broadband antennas, preferably without any termination resistors. As you will see in the following Sections, we have quite a number of options to choose from.

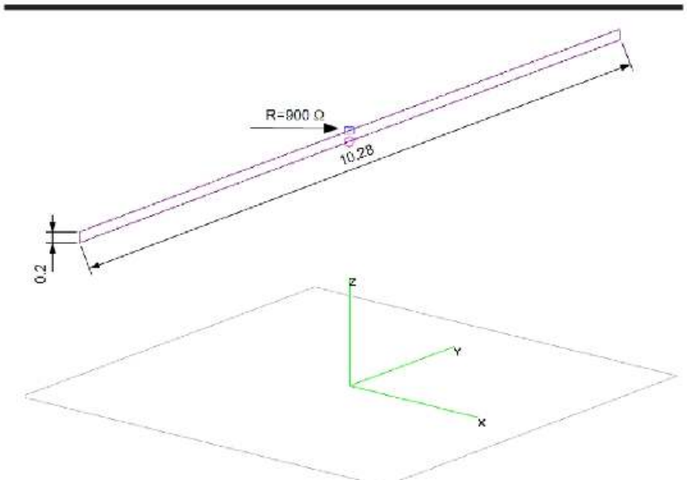

Figure 1 — Terminated Folded Dipole.

4 - Converging Element Doublets

Converging Element Doublet (CED) is the proposed name for the antenna in which a feed point of a straight wire element (dipole) is connected with a feed point of one or more companion V-shaped elements (V-dipoles) with the help of air insulated 2-wire line. The ends of the companion V-shaped elements come close to the ends of the basic dipole (converge to it) but are not electrically connected to them. See the figures below and this definition will become clearer.

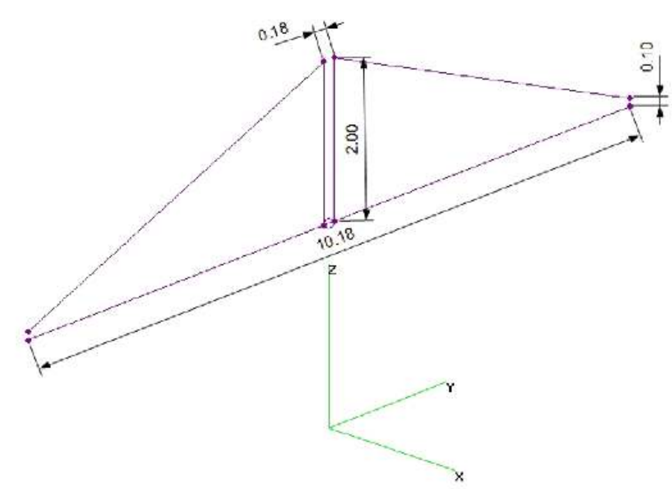

Figure 3 — Two-element CED is one of the simplest bidirectional broadband antenna. Its maximum radiation direction is along X-axis. All dimensions are in meters.

4.1 - Two-Element CED

Figure 3 shows the simplest two-element version of the CED. The straight element is 10.18 m long. The center points of the two elements are 2 m apart. They are connected with a 2-wire line having 0.18 m air gap. The ends of the companion element are 0.1 m away from the ends of the straight element.

The antenna has feed point impedance about 200 Ω, and it needs a 4:1 Guanella balun when fed with 50 Ω coax.

The plots in **Figure 4** (see **QEXfiles**) show SWR_{ANT}, gain and radiation patterns of the 2-element CED. The maximum value of SWR_{ANT} equals 3.9:1 at 22 MHz. The gain of the antenna itself at 14 MHz is almost exactly equal to the gain of a dipole, but if we take into account additional signal loss in the coax due to SWR, the relative gain is -1.2 dBdm. Less than a dipole, but this is a 5 dB improvement over the TFD!

At 29.7 MHz, the 2-element CED has gain +3.5 dBi. After correcting it down due to SWR, it is +0.2 dBdm. The doublet radiation lobes at that frequency are narrower than those of a dipole and there are no nulls at 90° and 270°.

4.2 - Three-Element CED

If we use not one but two companion elements, we will create a 3-element CED. The most obvious idea is to place all three elements in the same plane as shown in **Figure 5**.

Unfortunately, you can hardly see any improvement in performance over the previously presented 2-element CED. But if the outer elements are tilted up by some angle, we can get improvement in SWR_{ANT} and gain. Such an antenna is shown in

Figure 6. Its overall length is 10.3 m, distance between the wires in the air-insulated transmission lines connecting the elements is 0.5 m. Those transmission lines are 1 m long each and they are tilted 45° upwards. The ends of the elements converge down to 0.1 m. The feed point impedance is still about 200 Ω and a 4:1 Guanella current balun is required to feed it with 50 Ω coaxial cable.

The 3-element CED simulation results are shown in **Figures 7** and **8** (on **QEXfiles**) The SWR_{ANT} is equal to 3.4:1 at 14.0 MHz and 2.5:1 at 29.7 MHz. If we calculate the gain as explained before, we will get: -1 dBdm at 14.0 MHz and +1.1 dBdm at 29.7 MHz. The shape of the 3-element CED allows you to hide it under a roof after turning it upside down. But then you will have to find a way to rotate the whole building.

4.3 - Four-Element CED

The 4-element CED allows you to achieve even lower SWR_{ANT}. The antenna is presented in **Figure 9**. The 4-element CED dimensions are as follows: length is 11.44 m, separation of wires in the transmission lines is 0.48 m, vertical transmission line length is 0.9 m, horizontal transmission line lengths are 1 m each, gaps between the elements ends are 0.1 m.

Figures 10 and **11** (**QEXFiles**) show the performance graphs. A 4:1 Guanella balun is needed as before. Maximum SWR_{ANT} is only 2.7:1 and it happens at 29.7 MHz. This antenna is suitable for solid state transceivers without antenna tuning units. The comparison with a dipole looks like this: -0.6 dBdm at 14.0 MHz and +1.6 dBdm at 29.7 MHz.

4.4 - Five-Element CED

The 5-element CED allows you to achieve slightly lower SWR_{ANT} than a 4-element CED but all companion elements are now only 0.7 m away from the main element in the antenna center. This makes it a little easier to build. The antenna is shown in **Figure 12**. The 5-element CED dimensions are: length is 11.5 m, separation of wires in the transmission lines is 0.6 m, transmission lines lengths are 0.7 m each, gaps between the element ends are 0.1 m.

Figures 13 and **14** (**QEXFiles**) contain the performance graphs. Maximum SWR_{ANT} is only 2.6:1 and it happens at 14.0 MHz. This is also a perfect antenna for solid state transceivers without antenna tuning units. A 4:1 Guanella balun is needed as before.

If we compare this antenna with a dipole (both with feed lines), we will get the same results as for the 4-element CED, –0.6 dBdm at 14.0 MHz and +1.6 dBdm at 29.7 MHz. The radiation patterns look virtually the same as for the 4-element CED antenna.

4.5 - Which CED Is for You?

If you own a transceiver equipped with an antenna tuning unit (ATU), focus on the differences in dimensions and construction difficulty. With your ATU switched on, the difference in performance between different CEDs will be hardly noticeable.

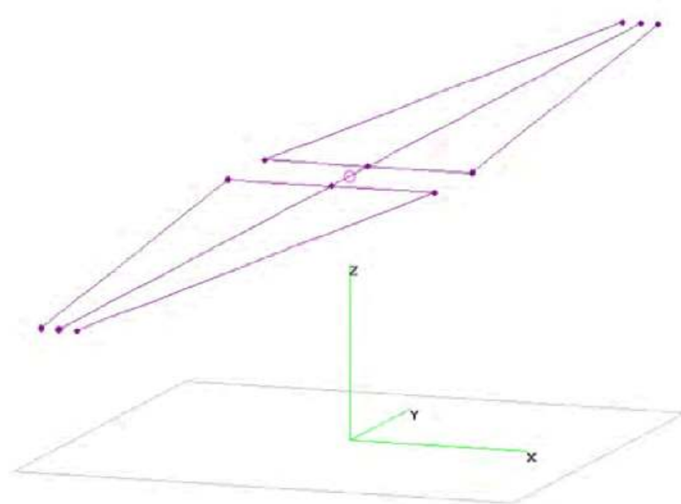

Figure 5 — One of the possible three-element CED designs. All three elements in the same plane.

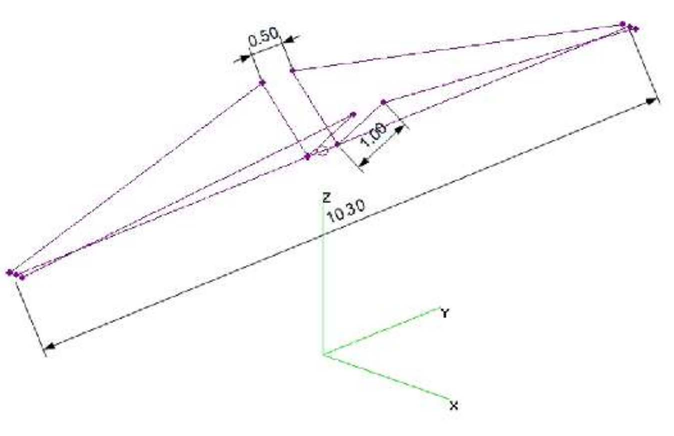

Figure 6 — This three-element CED shows improvement in SWR_{ANT} and gain over a two-element CED.

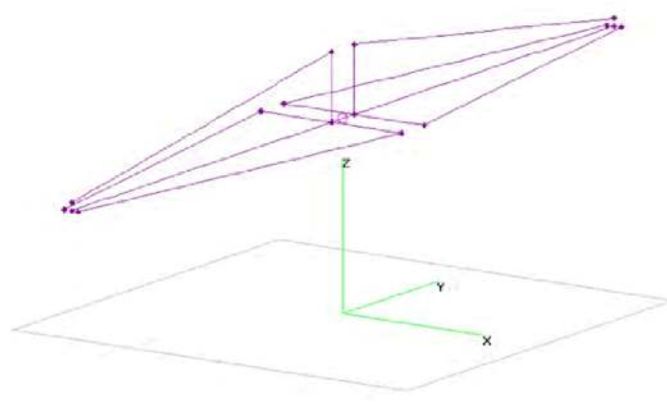

Figure 9 — Four-element CED configuration.

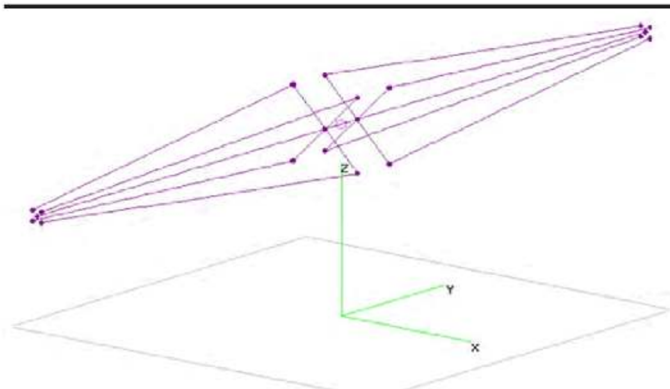

Figure 12 — Five-element CED configuration.

If you have a solid state transceiver without an ATU, you would better choose the 4-element or 5-element CED. You will need as low an SWR_{TRX} as possible. In this way, you will reduce power loss in the final stage of your transmitter.

Table 1 lists the dimensions and parameters of all versions of CEDs. Note that the *Radius* is the radius of an imaginary cylinder drawn around the main element in which the whole antenna fits.

5 - Spreading Element Doublets

After checking the idea of doublets with converging elements (CEDs), a logical next step is to try a doublet with elements that spread out rather than converge. **Figure 15** shows how the Spreading Element Doublets (SEDs) look. The SEDs can be oriented in horizontal or vertical plane. The difference in gain due to orientation is very small. Naturally, mounting them in horizontal plane is preferred because of a much smaller wind load.

The 5-element SED was additionally checked in a modified version (Mark 2) shown in **Figure 16**. The Five-element SED Mk.2 has a smaller element spread than the basic version, which perhaps makes it simpler to build. However, the elements are no longer in one plane, which increases wind load.

The plots of SWR_{ANT} and forward gain of the antennas without feed lines are shown in **Figure 17** (see **QEXfiles**) and the radiation patterns in **Figures 18**, **19** and **20** (see **QEXfiles**). **Table 2** compares four versions of SEDs.

As you can see in Table 2, the 5-element SED Mk.2 has the highest gain and almost the lowest SWR_{ANT} max. But its wind load is a little higher than the other versions, so the choice is not quite obvious. We have covered the doublets with converging and spreading elements. But can the elements be in parallel?

6 - Broadside Doublet Arrays

After modeling CEDs and SEDs, a time came to check if one can get a broadband antenna with parallel elements rather than converging or spreading ones. The elements are connected to one another with the help of an open-wire line as before.

An array of parallel dipoles equally arranged within one plane is called Broadside Dipole Array (BDA) or End-Fire Dipole Array (EFDA) depending on the direction of maximum radiation. The BDA and EFDA may look similar but in the case of BDA, the maximum radiation happens in a direction perpendicular to the plane in which the elements are arranged, and EFDA has strongest signal in a direction within this plane. **Figure 21** shows

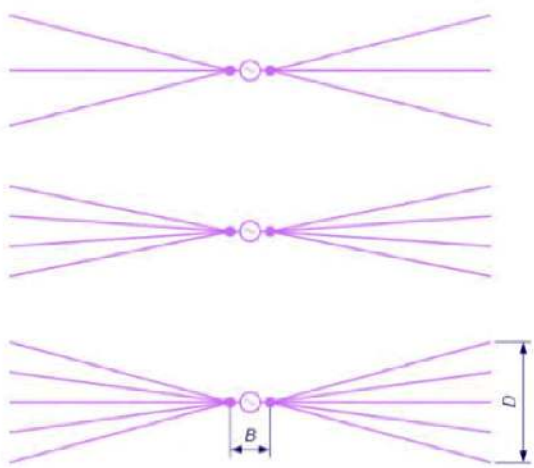

Figure 15 — Spreading Element Doublets (SED) from three-element (top) to 5-element (bottom) versions.

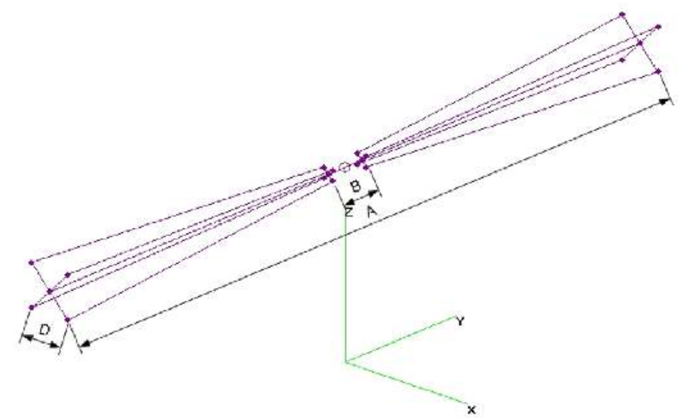

Figure 16 — Five-element SED Mk.2.

Table 1 – Comparison of different versions of CEDs and TFD.

Antenna type	*Length, m*	*Radius, m*	*max SWR_{ANT}*	*min Gain, dBdm*	*max Gain, dBdm*
2-ele CED	10.18	2.00	4.0:1	-1.2	0.2
3-ele CED	10.30	1.00	3.4:1	-1.0	1.1
4-ele CED	11.44	1.00	2.7:1	-0.6	1.6
5-ele CED	11.50	0.70	2.6:1	-0.6	1.6
TFD	10.28	0.20	2.5:1	-6.3	-3.3

Table 2 – Comparison of different versions of SEDs.

No. of elements	*A, m*	*B, m*	*D, m*	*Z_0, Ω*	*max SWR_{ANT}*	*min Gain, dBdm*	*max Gain, dBdm*
3	9.96	0.7	2.3	300	4.0:1	-1.3	-0.2
4	10.22	0.98	3.0	300	3.5:1	-1.0	0.8
5	10.10	1.2	1.5	300	3.9:1	-1.0	0.5
5 Mk.2	10.60	0.6	0.58	300	3.6:1	-0.9	1.6

the direction of maximum radiation for a 2-element BDA. For the best performance, the BDA should be oriented in the vertical plane. Such an upright position increases antenna wind loading and the tension it exerts on a rotator. If you position the BDA horizontally rather than vertically, it will lose some gain but will be less sensitive to wind. I did simulations for both positions — in vertical plane and in horizontal plane. *Gain V* refers to the gain when the BDA is in vertical plane and *Gain H* – when it is in horizontal plane.

Figure 22 shows what the BDAs look like. Note that the distance between the boundary elements is not necessarily increasing when the number of elements is increased. This is visible in **Table 3** where not only the dimensions of the BDAs are shown but also the main performance indicators. Some versions have their feed point impedance Z_0 closer to 200 Ω, while others are closer to 300 Ω. Depending on the version, you may need either a 4:1 or 6:1 current balun to feed it via 50 Ω coax.

One general remark. When you use an antenna optimizer program, your results depend on the initial values of *A*, *B* and *D* that you choose. It is quite possible, that one can find combinations of *A*, *B* and *D* producing equal or better results for a different feed-point impedance from those in the table. The results you can see are just the best that I got at the time of writing this text, but they are not necessarily the best possible in general.

Figure 23 (**QEXfiles**) shows the plots of SWR_{ANT} and gain for all four versions of BDA. The gain corrected due to SWR loss in the feed line is listed in **Table 3**.

That could be the end of the BDAs but I stretched my own rules concerning operational frequency range and maximum allowable SWR_{ANT} and tried to design BDA covering not five but six bands, that is, covering the frequency range from 10.1 MHz through 29.7 MHz. To achieve it, I needed to loosen the maximum SWR_{ANT} requirement to 5:1.

I found that for the 2-element and 3-element BDAs, I needed to make the antenna longer than 12 meters, which seemed to be too long. Fortunately, I succeeded with 4- and 5-element BDAs. Both BDAs were designed for 200 Ω source and required a 4:1 Guanella balun. They were both placed in horizontal plane. Their simulation

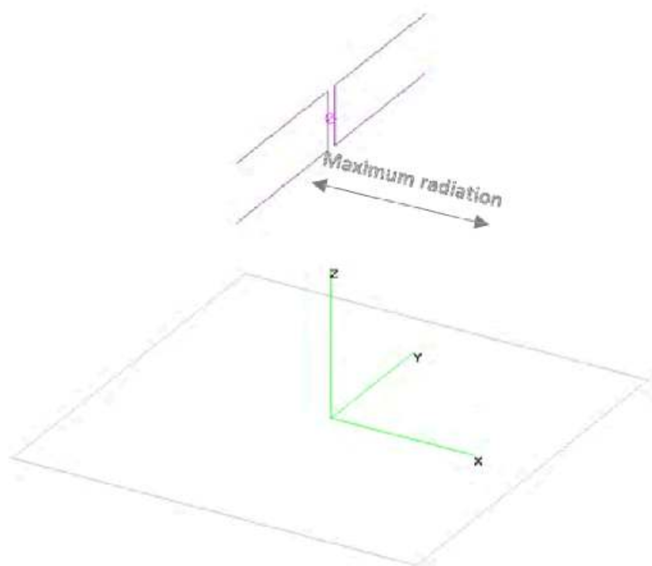

Figure 21 — Maximum radiation direction for a Broadside Doublet Array (BDA) occurs along the X-axis when the antenna is oriented in vertical plane.

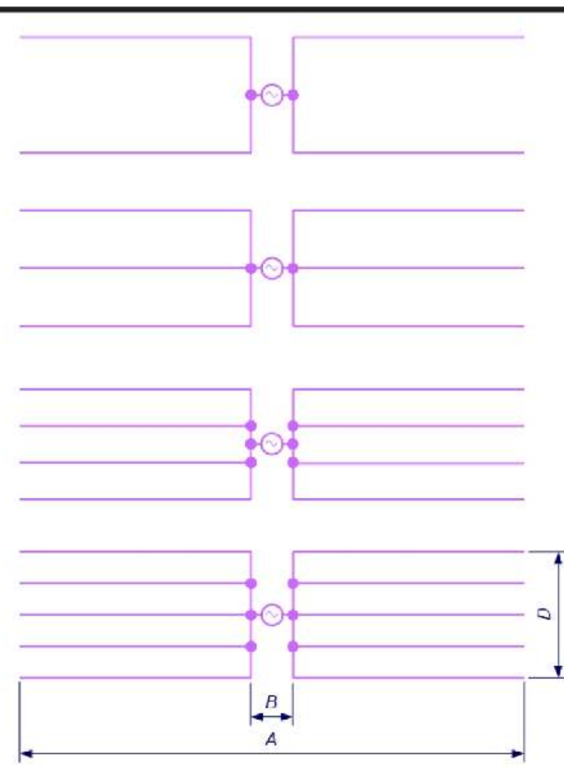

Figure 22 — BDAs from two-element (top) to five-element (bottom) version.

Table 3 – Comparison of different versions of BDAs.

No. of elements	*A, m*	*B, m*	*D, m*	*Z_0, Ω*	*max SWR_{ANT}*	*min Gain V, dBdm*	*max Gain V, dBdm*	*min Gain H, dBdm*	*max Gain H, dBdm*
2	11.46	0.42	2.00	300	3.9:1	-1.0	1.8	-1.4	-0.1
3	10.50	0.20	2.80	200	2.8:1	-0.5	2.1	-0.8	-0.4
4	10.64	0.38	1.74	200	2.8:1	-0.7	1.6	-0.7	0.8
5	11.14	0.44	1.60	300	3.1:1	-0.8	1.7	-0.8	1.0

Table 4 – BDA XFs (BDAs for extended frequency range: 10.1 – 29.7 MHz).

No. of elements	*A, m*	*B, m*	*D, m*	*Z_0, Ω*	*max SWR_{ANT}*	*min Gain H, dBdm*	*max Gain H, dBdm*
4	11.82	0.38	1.80	200	5.0:1	-1.9	2.1
5	11.80	0.40	1.72	200	4.8:1	-1.8	1.1

Table 5 – Comparison of different versions of EFDAs.

No. of elements	A, m	B, m	D, m	Z_0, Ω	max SWR_{ANT}	min Gain, dBdm	max Gain, dBdm
2	10.12	0.10	2.55	200	3.4:1	-0.2	2.5
3	10.60	0.16	1.94	200	2.9:1	-0.3	3.3
4	10.80	0.12	1.77	200	2.4:1	-0.2	3.3
5	10.64	0.20	1.68	200	2.4:1	-0.1	3.5

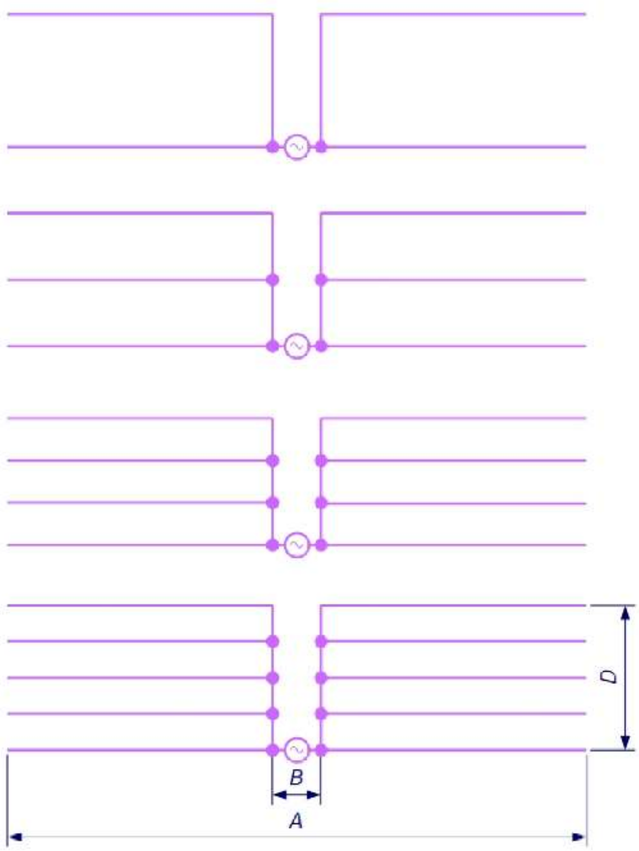

Figure 26 — End-Fire Doublet Arrays (EFDA) from two-element (top) to five-element (bottom) versions. Direction of maximum radiation is upwards in this drawing.

results are shown in **Table 4** and **Figures 24** and **25** (**QEXfiles**).

It is possible to create a broadband antenna covering 10.1 to 29.7 MHz on the basis of BDA. But you have to accept that at the lowest frequency, it will be 2 dB weaker than a dipole due to power loss in the coax. The 4-element version achieved higher gain than 5-element (by 1 dB). So, 4-element antenna seems to be optimum solution for a six-bander. However, because of its length — nearly 12 m — this antenna may be a challenge for less advanced antenna builders.

7 - End-Fire Doublet Arrays

The difference between EFDA and BDA is in the feed point location. In an EFDA, it is not located at the antenna center but at one of the outer elements. **Figure 26** shows EFDAs from 2-element through 5-element versions. Note that the maximum radiation for the end-fire array happens in its plane like in Yagis. So, it is normally positioned horizontally.

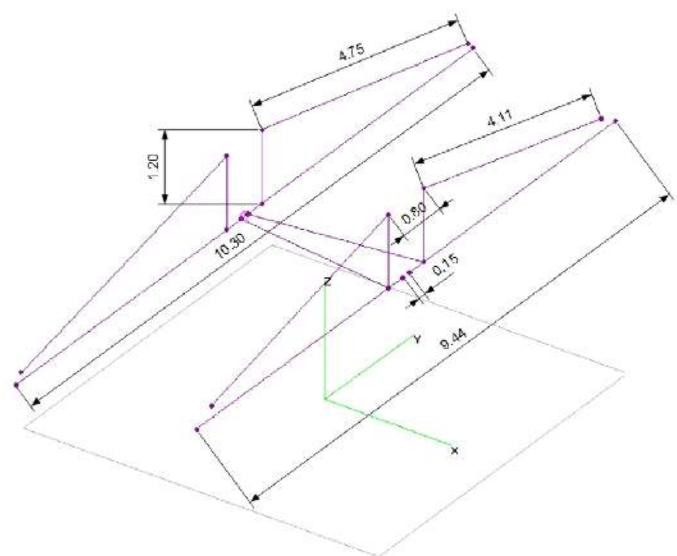

Figure 30 — Two-element EFDA Mk.2. Maximum radiation is towards +X.

EFDAs at the lowest frequency (14.0 MHz) have almost bidirectional radiation patterns. But they gradually turn into unidirectional patterns when the frequency is increased. This helps to achieve higher gains than for BDAs. **Table 5** list the dimensions and main performance parameters of different EFDAs. Both minimum and maximum gains are measured in the same direction but they happen at different frequencies. Minimum gain is around 14 MHz and maximum gain is around 29.7 MHz. Feed point impedance of all EFDAs is close to 200 Ω.

The plots of SWR_{ANT} and forward gain of the antennas without feed line are shown in **Figure 27** (**QEXfiles**) and their radiation patterns are in **Figures 28** and **29** (**QEXfiles**). The EFDAs feature higher gain and lower SWR_{ANT} than BDAs. At 14 MHz they are virtually equal to dipoles because -0.1 to -0.3 dB difference is almost nothing, and at higher frequencies they offer higher gain than a dipole. A gain of 3.3 through 3.5 dBdm in the 10 m band is remarkable. And even the simplest 2-element EFDA is surprisingly good. But can a good antenna be even better?

7.1 - Two-Element EFDA Mk. 2

Let's take 2-element EFDA and add converging elements to its two main elements. The new antenna view is seen in **Figure 30**. All antenna dimensions are shown in the figure except for a distance between the two elements, which is 2.55 m. The 2-element EFDA Mk.2 performance graphs are in **Figure 31** (**QEXfiles**). Note that the 2-element EFDA Mk.2 gain equals 2.7 dBi at 14 MHz and rises to 7.7 dBi at 29.7 MHz!

As you can see in **Figure 32** (**QEXfiles**), the 2-element EFDA

Mk.2 performs like a dipole on 20 m and like a 2-element Yagi on 10 m band (+4.8 dBdm). Its performance on the intermediate bands is between a bidirectional dipole and a unidirectional beam. That's what I call a real surprise!

[Part – 2 continues with Sections 8 through 10, including antenna detail Figures, and with performance details in the **QEXfiles** web page. — *Ed.*]

Jacek Pawlowski, SP3L, is an electronics engineer (M.Sc.). He started his professional career as an electronic designer, mainly in the test and measurement area. After about 15 years as a circuit/PCB designer, he switched to a management career path. He has been R&D project and department manager in a few companies since then. Jacek caught his radio bug when he was still in primary school in the early 1970s. In the years 1978-1999, he was active as SP3LFV. After that period, he suspended the hobby for 15 years. In 2014, he became active again under his present call sign and became interested in antenna design. This activity gives him a lot of satisfaction because it is like being a designer again.

Letter to the Editor

More Receiver Step Attenuator

I scratched my head several times while reading the "Receiver Step Attenuator" article by Scott Roleson, KC7CJ, *QEX* Jul./Aug. 2019. I would like to offer my own alternative perspectives that might benefit readers. I built my first attenuator around 1985 for transmitter hunting. It was not a precision instrument, but provided continuous adjustment over a 60 dB range. It consisted of two 500 Ω potentiometers. They were built inside a box made from PCB material, and included a partition separating them. The biggest problem I encountered was the grounding of the bottom side of the potentiometers. Even a quarter inch of wire between the potentiometer lug and the chassis was enough inductance to degrade the maximum attenuation at 148 MHz. Lesson learned.

The attenuator in the article was intended to provide only 21 dB of attenuation, which should be very easy to achieve. I'm not sure that partitions were even necessary.

My first suggestion would be to lay the circuit out with one side of the PCB as a solid ground plane. That could make a huge improvement. Once that is done, the traces that involve the RF should have their widths calculated to be consistent with 50 Ω microstrip, so no resonances or reflections are present.

The second suggestion would be to use Surface-Mount Technology (SMT) resistors. They have a much lower parasitic inductance and much more closely approximate the 50 Ω impedance of the traces. Making those two changes would probably result in a design that could easily achieve expected performance through the VHF range, even without divider shields.

In order for divider shields to be effective they must provide 100% separation. Normally the enclosure is hogged out of a block of aluminum or brass with a milling machine, with partitions in place, and has a matching cover that bolts tightly to it with about 20 screws. Soldering metal flags to a PCB is not a very close approximation to this. The poor man's version is to use un-etched, double-sided PCB material (or metal sheets) to form the enclosure and dividers, and solder the dividers around all edges, so there are no gaps.

The point of using relays, while obviously a designer's preference, eluded me. It would seem that the traditional and often published use of slide switches would be equally useful, possibly with lower leakage, and would be less costly. That is unless, off course, there was a need to remotely control or automate the attenuator.

Finally, given today's technology, there is an even simpler and better approach to this problem. Several companies make attenuator ICs with ranges to over 30 dB in a single IC and step-sized as small as 0.25 dB. Some are very accurate and cover the range from dc to more than 1 GHz. My present favorite is the PE4312 digital step attenuator, which covers a 31.5 dB attenuation range in 0.5 dB steps and operates from 1 MHz through 4 GHz. It can be controlled either by binary-weighted control lines or SPI serial interface. It offers 1% accuracy up to 1 GHz. It does require power, but so do relays. It can be powered from 2.3 to 5.5 V and consumes less than 1 mA. A couple of flashlight cells would probably last nearly their shelf life if one wanted to do it that way. Furthermore, the cost is under $5.00 — less than the relays in the KC7CJ design.

So what's the downside? The PE4312 is a relatively small SMT IC, and many hams are intimidated by SMT designs. That is unfortunate, because many current IC designs (especially RF ones) use SMT. Our hobby is primarily about RF, and SMT devices exhibit behavior much closer to ideal than do designs using through-hole parts. It may require having a company etch the PCB. Also, plated-through holes are difficult to do or emulate at home, and are essential to a quality RF board design. As for assembly, these ICs pose two challenges. The first is that there is a ground pad in the middle of the underside that must be soldered down. Second, the 'pins' are pieces of metal on the under side that extend just to the edge of the package. While the latter can be soldered with a fine iron, especially if the pads extend beyond them, the easiest way to solder one of these without special equipment is to lightly tin both the contacts on the bottom of the IC and the pads they are supposed to attach to, then position the IC in place and gently heat it and the board with a heat gun until the solder flows and it seats itself appropriately. An alternative approach would be to place the board in a toaster oven, but the slow response to temperature changes that is characteristic of the oven could require some experimentation to get it right.

Yes, it's a bit messier for a one time project, but the simplicity of the design and the potential improvement in performance is well worth it. I recently designed a piece of test equipment using five of these SMT attenuators to provide 140 dB of adjustment range. Yes, the last three were in individually shielded compartments, to prevent stray signals from leaking in. — *Best regards, Wilton C. Helm, WT6C,* **whelm@compuserve.com**.

Anchors for a Printed Lid

In the July/August 2019 issue of *QEX*, Joseph Pingree, WB2TVB, in "Get Started with 3D Printing" describes the utility of *OpenSCAD* for the design and fabrication of 3D-printed hardware. I have also used this excellent, programmer-oriented software package to render a variety of custom components including small enclosures. In Figure 3 of the article, a method for securing the enclosure lid can be found as a comment in the design code. Cylindrical holes are placed in plastic pillars at each interior corner that are then reamed and tapped to produce the threads.

In my designs, I use threaded brass anchors that can be obtained for low cost on eBay. The anchors are press-fit into holes in each pillar using the tip of a heated soldering iron. This takes only a few seconds and leaves a robust anchor that is melted into the plastic (see **Photo 1**). The lid can then be firmly screwed into place with little concern for stripping delicate machine-screw threads.

The brass anchor should be inserted into a tapered hole in the plastic. The recommended taper can be readily implemented with a few lines of code in *OpenSCAD*. — *Best regards, Mike Hasselbeck, WB2FKO,* **mph@sportscliche.com**

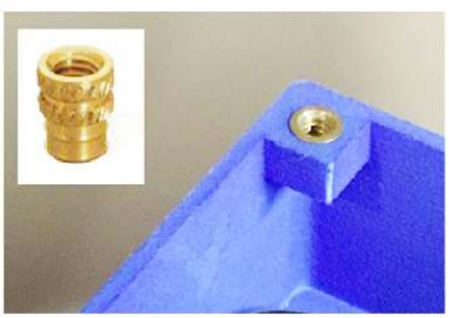

Photo 1 — Brass anchor is inserted into a tapered hole in the plastic.

Small HF Loop Antennas are not Magnetic Loops

Small HF loop antennas, typically 1 m in diameter, have been mistakenly called "magnetic loops." They are not, and here's why. From the solution to Maxwell's equations, the current around the circumference ϕ of the small loop is a Fourier series in terms of $\cos(n\phi)$. Thus there is a current variation around the loop circumference. Using just two ($n = 0$ and $n = 1$) Fourier series terms is sufficient to accurately describe the current around the typical small HF loop. The amplitude of the $n = 1$ term equals twice the square of the loop circumference in wavelengths in ratio to the $n = 0$ constant current term thus depends only on C_λ, the loop circumference in wavelengths. The $n = 1$ current variation term produces a near electric-to-magnetic-field ratio (wave impedance) at the loop center that is proportional to C_λ multiplied by the free space intrinsic impedance. Furthermore, the current variation term produces a far-field component that fills in the loop null, resulting in a null depth that is the square of twice C_λ (and is usually expressed in decibels). The exact solution to Maxwell's equations for the loop current, taking into account just the first two ($n = 0$ and $n = 1$) Fourier terms, from Eqn (3) in reference **[1]** is,

$$I(\phi) = I_0\left\{1 - 2C_\lambda^{\,2}\cos(\phi)\right\}$$

where ϕ is the angle around the loop circumference starting at the loop gap, and I_0 is a constant related to the power transmitted by the loop. **Figure 2 in [1]** repeated here for convenience, illustrates the loop current variation across 7 MHz to 30 MHz.

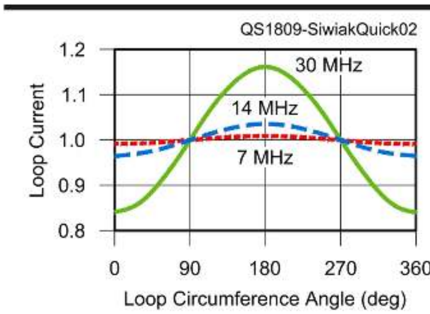

Figure 2 in [1] — Loop currents along the loop circumference at 7, 14, and 30 MHz.

The corresponding Eqn (22) in [1] for the loop near-field wave impedance or near-electric to magnetic fields ratio at the center of the loop is,

$$Z_W = \frac{E_{center}}{H_{center}} = -j\eta_0 C_\lambda$$

where $\eta_0 = 376.73\ \Omega$ is the free space intrinsic impedance. Eqn (23) in [1] for the null depth in dB is,

$$N_{dB} = -20\log(2C_\lambda).$$

These equations are *exact solutions* for two Fourier terms and depend only on the loop circumference C_λ. They are accurate for a loop circumference of up to about 0.3 or 0.4 wavelengths before additional ($n > 1$) Fourier loop current terms are needed. As the loop circumference grows, the $n = 1$ and higher order terms gradually begin to dominate, and at $C_\lambda = 1$ we have the well known one-wavelength diameter loop that radiates broadside to the loop plane, in the direction that is the null direction for $C_\lambda << 1$. These results can be easily verified and validated using NEC software, such as *4nec2* or *EZNEC*, as in [1]. Note that the field in the broadside direction is cross-polarized relative to the field in the loop plane.

Now we have three *exact* equations that depend solely on the loop circumference in wavelengths: one for the current, one for the near-field wave impedance, and one for the null depth. Thus we have three inter-dependent ways of choosing *objective criteria* for when a loop can qualify as a "magnetic-field loop" (not magnetic loop) based on the loop size. We can choose one of these three criteria: {1} define an acceptable far-field null depth, or equivalently {2} we could define the maximum acceptable near-field electric-to-magnetic-field ratio (the wave impedance at the loop center), or we can {3} define the maximum acceptable loop current variation. A true ideal "magnetic-field loop" or a "magnetic-field probe" would have {1} an infinitely deep null, and {2} a wave impedance or near electric-to-magnetic-field ratio of zero ohms, and {3} constant loop current, implying a vanishingly small loop circumference.

Some references arbitrarily define a small loop as having a loop circumference of under 0.2 wavelengths. However, that results in a null depth of just 8 dB — hardly infinite! — and a near electric-to-magnetic-field ratio (wave impedance) of 75 Ω — hardly close to zero! — and a loop current variation of ±8% — hardly a constant current!

F. M. Greene, see **[2]**, suggests that the diameter of a magnetic-field probe should not exceed $\lambda/100$ ($C_\lambda = 0.031$) resulting in a null depth of 24 dB and near electric-to-magnetic-field ratio of 3.1% of the free-space intrinsic impedance. I propose a slightly looser definition of a "magnetic-field loop" antenna as having a far-field null that is at least 20 dB deep, which requires the loop circumference to not exceed $C_\lambda = 0.05$ wavelengths. The corresponding near electric-to-magnetic-field ratio would be 5% of the free-space intrinsic impedance (about 19 Ω), and the loop current variation would be less than ±0.5%. Typical 1 m diameter loop antennas would meet these "magnetic-field loop" criteria at frequencies below 4.8 MHz. Since the operating range of 1 m diameter small HF loops is between 7 MHz and 29.7 MHz, they are *never* magnetic-field loops.— *Kindest regards, Kai Siwiak, KE4PT,* **k.siwiak@ieee.org**

Notes

[1] K. Siwiak, KE4PT, and R. Quick, W4RQ, "Small Gap-resonated HF Loop Antenna Fed by a Secondary Loop", *QEX*, Jul./Aug. 2018, pp. 12-17: online at **www.arrl.org/files/file/QST%20Binaries/September2018/QSTinDepth-Sept-2018-Siwiak-Quick.zip**.

[2] F. M. Greene, "Development of Electric and Magnetic Near-Field Probes," National Bureau of Standards Report No. NBS TN-658, Jan. 1975, p. 46.

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Upcoming Conferences

SCALE 18x

March 5 – 8, 2020
Pasadena, California
www.socallinuxexpo.org/scale/18x

SCALE 18x, the 18th annual Southern California Linux Expo, will take place March 5 – 8, 2020, at the Pasadena Convention Center. SCALE 18x expects to host 150 exhibitors this year, along with nearly 130 sessions, tutorials and special events.

SCALE is the largest community-run open-source and free software conference in North America. It is held annually in the greater Los Angeles area.

2020 SARA Western Conference

March 27 – 29, 2020
Socorro, New Mexico
www.radio-astronomy.org

The 2020 SARA Western Conference will be held at the Pete V. Domenici Science Operations Center in Socorro, New Mexico, March 27 – 29, 2020.

The keynote speaker will be Dr. Mark McKinnon, Assistant Director for New Mexico Operations at National Radio Astronomy Observatory. In addition to presentations by SARA members, plans include having other speakers from the NRAO SOC in Socorro, and from the UNM Long Wavelength Array (LWA) project. On Sunday, March 29, we will have a tour of the Very Large Array (VLA) site west of Socorro. Additional details will be published online and in the SARA journal as we get closer to the conference date. Register now to avoid the rush and to guarantee a seat at the conference.

Registration: Registration for the 2020 Western Conference is just $80.00. Attendees at the conference must be SARA members; if you are not yet a member, this will cost an additional $20. The fee includes lunch and snacks on Friday and Saturday and lunch on Sunday. See website for payment options.

Hotel reservations: Reservations can be made at the Best Western Hotel & Suites in Socorro by calling (575) 838-0556. Be sure to tell them that you will be attending the SARA Western Conference. The special conference rate is $131.99 per night (109.99 + tax per night single or double occupancy).

2020 Southeastern VHF Society Conference

April 24 – 25, 2020
Gainesville, Georgia
www.svhfs.org

The 2020 Southeastern VHF Society Conference will be held April 24 – 25, 2020 at the Ramada Inn in Gainesville, GA. Hotel reservations are currently being accepted; the conference rate is $80 per night. To secure this rate, call the hotel directly at 770-531-0907. Do not call Ramada central reservations or book through the Ramada website.

Papers and presentations are solicited on both the technical and operational aspects of VHF, UHF and Microwave weak signal amateur radio. The deadline for the submission of papers and presentations for this year's conference is March 7, 2020. Some suggested areas of interest are: transmitters; receivers; transverters; RF power amplifiers; RF low noise preamplifiers; antennas; construction projects; test equipment and station accessories; station design and construction; contesting; roving; DXpeditions; EME; propagation (sporadic E, meteor scatter, troposphere ducting, etc.); digital modes (WSJT, etc.); digital signal processing (DSP); software defined radio (SDR); amateur satellites, and amateur television.

In general, papers and presentations on non-weak signal related topics such as FM repeaters and packet will not be accepted but exceptions may be made if the topic is related to weak signal. For example, a paper or presentation on the use of FM simplex in contests or on the use of APRS to track rovers during contests would be considered. Papers and presentations are being handled by Charles Osborne, K4CSO, and should be sent to: **k4cso@twc.com**.

A current conference schedule and other information is available on the website.

Aurora Conference

April 25, 2020
White Bear Lake, Minnesota
www.nlrs.org/home/aurora

The 2020 Aurora conference will be held April 25 at the Community of Grace Lutheran Church, 4000 Linden St., White Bear Lake, MN. Aurora, the largest annual gathering of weak-signal VHF'ers in the Upper Midwest, is the annual gathering of the Northern Lights Radio Society. These conferences, first hosted by Jon Lieberg, KØFQA, began in 1984.

In the morning hours, the club runs an outdoor antenna range from 0900 to 1130. The morning also allows for socializing, show & tell, and casual demonstrations. Members are then on their own for lunch. The technical programs start at 1300 and typically run until 1700 – 1730.

If you have a weak signal VHF topic that is of interest to you, or that you would like to present on, please contact our Technical Chairman, Jon Platt, WØZQ (**w0zq@aol.com**). If a technical presentation is not to your liking, Aurora also features a poster session.

If you have an interest in weak signal VHF please plan on attending Aurora where you will leave re-energized for this exciting aspect of our hobby. Watch website for details.

Central States VHF Society Conference

July 24-25, 2020
La Crosse, Wisconsin
2020.csvhfs.org

The 54th annual Central States VHF Society Conference will be held at the Radisson Hotel located on the beautiful riverfront of the Mississippi River in La Crosse, Wisconsin on July 24 – 25, 2020.

This year's event will have all the great activities you've come to expect from a CSVHFS Conference: technical presentations, antenna range, noise figure lab, rover row and dish bowl, Thursday evening social activity, Friday evening trade-fest, dealer room, hospitality suite for evening socializing, fun family activities, and a closing banquet with a guest speaker and a prize table.

Papers are being solicited for publishing in the Proceedings of the 2020 Central States VHF Conference on all weak-signal VHF and above amateur radio topics, including: antennas (modeling, design, arrays, and control); test equipment (homebrew, commercial, and measurement techniques and tips); construction of equipment such as transmitters, receivers, and transverters; operating (contesting, roving, and DXpeditions); RF power amps (single and multi-band vacuum tubes, solid-state, and TWTAs); propagation (ducting, sporadic E, tropospheric, meteor scatter, etc.); preamplifiers (low noise); digital modes such as WSJT, JT65, FT8, JT6M, ISCAT, etc.; regulatory topics; moon bounce (EME); software-defined radio (SDR) and, digital signal processing (DSP).

Topics such as FM, repeaters, packet radio, etc., are generally considered outside of the scope of papers being sought. However, there are always exceptions. If you have any questions about the suitability of a particular topic, please contact Kent Britain, WA5VJB, **wa5vjb@wa5vjb.com**, or Donn Baker, WA2VO, **lwa2voi@mninter.net**.

You do not need to attend the conference nor present your paper to have it published in the Proceedings. Posters will be displayed during the Conference.

Reservations for rooms will open in February. Check website for a hotlink that will take you directly to the Radisson website for event room rate reservations.

The negotiated room rates are available from July 20 through July 28. If you choose to arrive early or extend your stay, these rates will be available to you.

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