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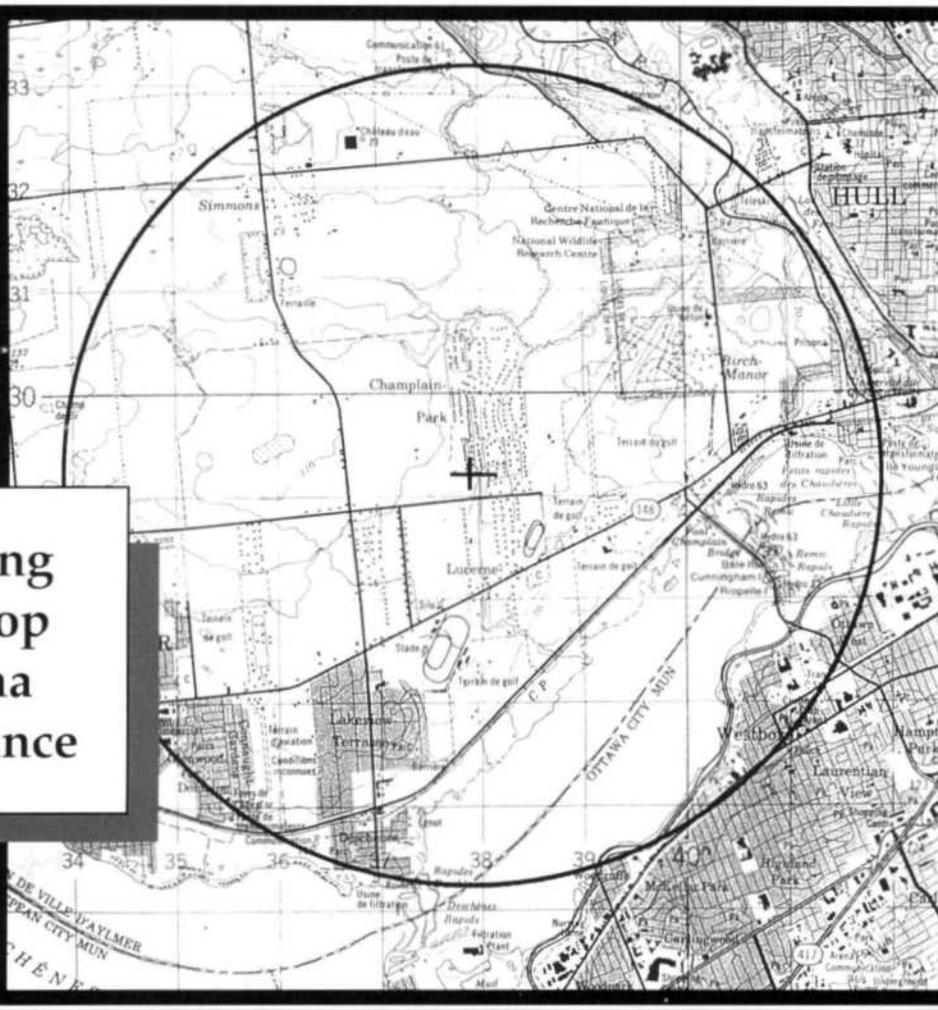
ARRL Experimenter's Exchange

August 1997



Analyzing Real Loop Antenna Performance

QEX: The ARRL
Experimenter's Exchange
American Radio Relay League
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David Sumner, K1ZZ
Publisher

Rudy Severns, N6LF
Editor

Robert Schetgen, KU7G
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Production Assistant

Advertising Information Contact:

Brad Thomas, KC1EX, *Advertising Manager*
American Radio Relay League
860-594-0207 direct
860-594-0200 ARRL
860-594-0259 fax

Circulation Department

Debra Jahnke, *Manager*
Kathy Capodicasa, N1GZO, *Deputy Manager*
Cathy Stepina, *QEX Circulation*

Offices

225 Main St, Newington, CT 06111-1494 USA
Telephone: 860-594-0200
Telex: 650215-5052 MCI
Fax: 860-594-0259 (24 hour direct line)
Electronic Mail: MCIMAILID: 215-5052
Internet:qex@arrl.org

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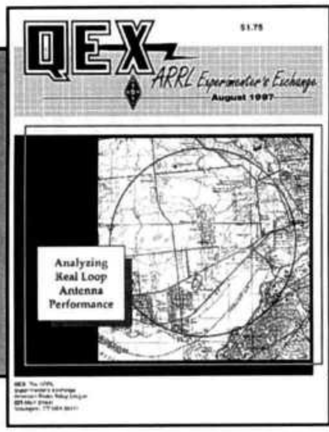

About the Cover

VE2CV shows how several kinds of large HF loop antennas perform using analytical tools.

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THE AMERICAN RADIO RELAY LEAGUE

The American Radio Relay League, Inc, is a noncommercial association of radio amateurs, organized for the promotion of interests in Amateur Radio communication and experimentation, for the establishment of networks to provide communications in the event of disasters or other emergencies, for the advancement of radio art and of the public welfare, for the representation of the radio amateur in legislative matters, and for the maintenance of fraternalism and a high standard of conduct.

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1) provide a medium for the exchange of ideas and information between Amateur Radio experimenters
2) document advanced technical work in the Amateur Radio field
3) support efforts to advance the state of the Amateur Radio art

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

Both theoretical and practical technical articles are welcomed. Manuscripts should be typed and doubled spaced. Please use the standard ARRL abbreviations found in recent editions of *The ARRL Handbook*. Photos should be glossy, black and white positive prints of good definition and contrast, and should be the same size or larger than the size that is to appear in *QEX*.

Any opinions expressed in *QEX* are those of the authors, not necessarily those of the editor or the League. While we attempt to ensure that all articles are technically valid, authors are expected to defend their own material. Products mentioned in the text are included for your information; no endorsement is implied. The information is believed to be correct, but readers are cautioned to verify availability of the product before sending money to the vendor.

Empirically Speaking

A New Editor

Now that *QEX* is back, we have made some changes to keep it on track. Jon Bloom has been the *QEX* editor for five years and did a great job. We all owe him our thanks. As we pointed out in July *QEX*, one of the problems we have had is that the editor and support people at headquarters had to serve more than one master, and Jon was pulled off to do other, albeit important, tasks. One solution is to appoint an "outsider" who is not so easily distracted by other priorities at headquarters. That's where I come in. I am Rudy Severns, N6LF, and I have volunteered to take over as editor of *QEX*. Many years ago, in different circumstances, I volunteered and found myself jumping out of airplanes in the middle of the night. I swore never to volunteer again, but memory dims and we forget our resolutions. Besides, I feel I have gained much from Amateur Radio and perhaps it's time to pay back a little. I was first licensed as WN7WAG in 1953 and have been continuously licensed since then. I have always felt that ham radio was an important part of my life and it certainly has helped me greatly both in my military service and later in my professional career as an electrical engineer.

I, and I am sure most of you, have noticed (and lamented) the steady decline in technical content in amateur publications. Some of this decline is understandably due to changes in our hobby, but I have never accepted that we are, or should become, nothing more than appliance operators. That is not in keeping with the tradition I know. I view *QEX* as an important vehicle for keeping our technical tradition alive.

It is my view that *QEX* should contain technical articles at many different levels: from the simple construction article to the technically complex. Hopefully, each issue will be a mix of levels and subjects. There are also subjects which are controversial, on which we are not all in agreement. As long as the discussion is civil and constructive, I think *QEX* is an appropriate forum. A forum which really doesn't exist anywhere else. As editor, my job is to act as referee, cheerleader, bookkeeper, scheduler and crying shoulder. While I will occasionally contribute articles of my own, most of the material must come from you, the readers of *QEX*. If you don't see what you want in *QEX*, write it! Or at least ask for it. Until that far distant day when I am snowed under with wonderful articles, you will find me very accepting of new material. While some reasonable standard for clarity and accuracy must be applied, we are not creating an IEEE transaction. Remember, we do this for fun. Don't worry about perfection in writing style. Put your work in writing as best you can and share it with all of us. I have neither the time nor the expertise to judge every submission. We are fortunate enough to have a number of very qualified technical advisors to whom I can turn for review and comment on submissions.

Enough said, now it's time to get to work.

This Month in *QEX*

DX on 80 meters. It's one of those things that you're likely to gravitate toward once you've decided that the upper bands are "too easy." And, as on those upper bands, maybe even more so, it's the antenna that makes the difference. "Loops for 80-Meter DX," by John S. Belrose, VE2CV, explores one set of choices of 80-meter DX antenna. It may be a choice that makes sense for you. Follow along as Jack analyzes the performance of 80-meter loops of various configurations.

Many hams are interested in older, tube-type equipment. Whether it's due to nostalgia or simply an interest in wringing the most out of older technology (or wringing the most out of your ham-radio dollar), tube-type equipment still moves at the hamfest flea-market table. But getting replacement tubes, even common types, at reasonable prices is getting harder and harder. So, Parker R. Cope, W2GOM/7, shows how "Synthesizing Vacuum Tubes" using FETs can bring life back to those old tube receivers.

Finally, "Upcoming Technical Conferences" once again points you to where the leading edge of Amateur Radio is happening.—*73, Rudy Severns, N6LF, n6lf@arrl.org*

Loops for 80-Meter DX

VE2CV has been experimenting with and modeling low-frequency antennas for many years. Jack recounts his extensive practical experience using different antennas on the lower bands.

By John S. Belrose, VE2CV

A simple yet effective antenna for the 80 and 160-meter bands is the half-wave dipole, although other antennas are used, particularly by those interested in working DX. Propagation conditions during nighttime and evening hours support communication on both long and short paths. The station antenna or antennas ideally should radiate equally well at both low and high radiation angles, also referred to as *elevation angles* (ψ) measured from the Earth's surface.

For communication with a distant station, you must listen to the signal against a background of noise and interference, particularly interference from strong, near-vertical-incidence skywave (NVIS) signals ($\psi > 60°$). In addition, local man-made noise and atmospheric noise from nearby thunderstorms is almost always present. Horizontal polarization is preferred over vertical polarization from the ground reinforcement point of view, unless you are fortunate to have seawater in front of the antenna in the desired direction(s). But practical dipole heights for the 80-meter band are too low to achieve an effective low angle of radiation ($\psi < 15°$). This is even more of a problem for the 160-meter band.

Since I became a radio amateur, I have been interested in devising efficient antennas that resolve some of these difficulties. While I'm not a DXer, from my QTH in Aylmer, QC, I like to "check into" distant 80-meter nets, such as the North West Ontario Net or the Newfoundland Phone Net. For almost a decade, I've been a control station for the Trans-Canada Pow-Wow Club, a group of amateurs that meets on-the-air daily during winter and equinox months, beginning at midnight local Eastern time on 3750 kHz. I am the control station for the Sunday morning club meetings.

Pow-Wow Club members and other amateurs attracted by the activity on the frequency call in from across Canada, coast-to-coast, as well as from the UK. Participation in the Club is intended to inspire members to improve their stations' capabilities to a point where they can be heard from coast-to-coast. Linear amplifiers are acquired or constructed and antennas are put up, modified and put up

ARRL Technical Adviser
17 Tadoussac Drive
Aylmer, QC J9J 1G1 Canada

again—antennas that radiate as well as possible at both low and high elevation angles. Beverage antennas are used by some to improve the received signal-to-noise. I use three simple antennas:

- a dipole—for many years 15-meters high, but presently at 26 meters, the height of the trees,
- a ground-plane type of delta loop, an antenna that I devised and dubbed a *half-loop*[1],
- a compact loop, an AMA-11, a 1.7-meter diameter loop used primarily for receiving[2].

These antennas have very different radiation patterns, and their different responses to noise and interference is a distinct advantage. The ability to switch quickly between two (or more) antennas is important, since this provides some ability to optimize reception in the face of variable propagation conditions, noise and interference.

Half-Loop Antennas

For a decade and a half I have employed loops in various configurations, but the loop antenna I've used for the longest time is the *half-delta loop*. This vertically polarized loop has a perimeter of only half an electrical wavelength, since the other half of this ground-mounted loop is its image in the ground plane. An article by Al Christman, N3AL (ex-KB8I), and follow-on studies by me (using the Numerical Electromagnetics Code, *NEC-2*) show that the radiation efficiency for a monopole with only three or four resonant elevated radials is equivalent to that for a grounded vertical monopole with 120 buried radials.[3,4] The result of this work inspired me to consider lifting my half loop off the ground, using radials to simulate connection to ground.

The Effect of Real Ground

First, we need a brief discussion of the effect of finitely conducting ground on antenna performance. After that we will consider the ground-mounted half-wave loop in detail. In the case studies that follow, I created radiation patterns using *EZNEC*, developed by Roy Lewallen, W7EL.[5] *EZNEC* is a menu-driven version of *NEC2*, the Numerical Electromagnetics Code. Unless otherwise stated, the frequency is 3.75 MHz and the patterns have been calculated for resonant loops. When I refer to "principal-plane patterns," this means the azimuthal pattern at the elevation angle of maximum gain, and the elevation pattern for the azimuth angle of maximum gain.

Both directly beneath and in front of an antenna in the direction of propagation, lossy ground affects the antenna's impedance, the current on the antenna and its radiation pattern. If ground conductivity is poor, it is better to use a horizontally polarized antenna. The gain of a horizontally polarized antenna, particularly at low launch angles, is less affected by the finite conductivity of the ground in front of the antenna.

Ground reinforcement of vertically polarized waves results in a pattern maximum at the angle of launch (ψ_{max}), where the direct and the ground-reflected rays are in phase. This occurs at low launch angles for "perfect" ground. But when the ground conductivity is finite, the phase of the ground-reflected ray changes rather abruptly from an in-phase to an out-of-phase condition at a particular launch angle called the *Brewster angle*. For launch angles below the Brewster angle, antenna performance is seriously degraded.

Over seawater this happens at launch angles that are a fraction of a degree above the horizon, and the formation of the low-angle maximum is scarcely affected. Over typical soil, the phase change occurs at launch angles around 15°. For launch angles below this, performance is seriously degraded. (See Fig 1A.) This poor performance occurs at angles comparable to the most probable arrival angles for skywaves from distant stations. [See Chapters 3 and 23 of the 18th edition of *The ARRL Antenna Book* for detailed elevation-angle statistics.—*Ed.*]

For horizontal polarization over imperfect ground, the reflection coefficient at small elevation angles is slightly smaller than at higher angles. The phase change for the ground-reflected wave is substantially 180°, so the launch angle for maximum radiation is scarcely changed. The reduced reflection coefficient leads to a small loss of gain, by a dB or so for low launch angles (see Fig 1C). For high launch angles, the ground-reflected wave is weaker and the

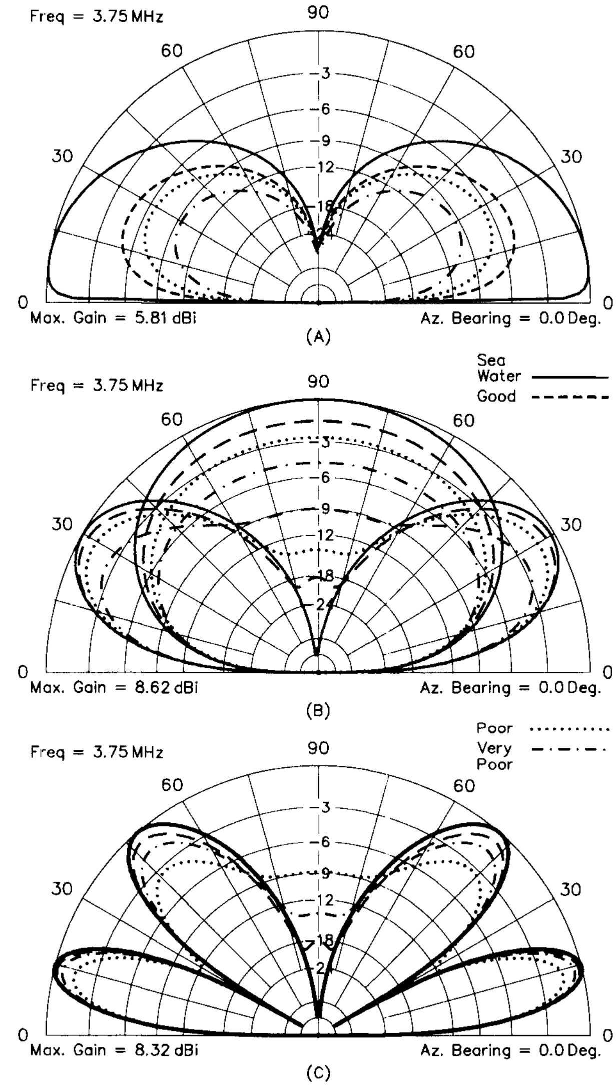

Fig 1—At A, elevation pattern for half-diamond GP-type loop. At B, elevation pattern for a dipole at 15 meters (0.19 λ) and at 40 meters (λ/2). At C, elevation pattern for a dipole at 80 meters (1 λ). Frequency in all cases is 3.75 MHz, for four ground conductivities: sea water (solid line), good ground (dashed line: σ = 10 mS/m, ε = 30), poor ground (dotted line: σ = 1 mS/m, ε = 15), and very poor ground (dashed-dotted line: σ = 0.1 mS/m, ε = 3).

[1]Notes appear on page 16.

antenna's gain is less, by up to 4 dB compared to perfect ground. The gain does remain some 2 to 3 dB higher than it would be if ground reinforcement were absent. (See Fig 1B.) For practical purposes, and particularly for antennas designed for low launch angles, the effects of real ground may often be ignored if the polarization is horizontal.

The launch angle for maximum gain from a horizontal dipole, however, depends on the height in wavelengths of the antenna above the ground. For dipole heights less than 0.3 λ, ground reinforcement occurs at high launch angles; that is, ψ close to 90°. This is an antenna that Doug DeMaw, W1FB, refers to as a "cloud warmer." (See Fig 2A.) When the dipole height is 0.5 λ, the radiation pattern has a single lobe in the plane orthogonal to the plane of the dipole, and there is a null overhead. For ground reinforcement to occur at small launch angles, say $\psi < 15°$, the electrical height of the dipole must be greater than one wavelength. For 80 meters, this is impractical, since the height would have to be greater than 80 meters! The best antenna (vertical or horizontal polarization) for 80-meter DX is obviously a compromise.

Ground-Plane Type Half-Loops

Full-wavelength loops have been popular with radio amateurs for many years. They are used as elements for quad and delta-loop beam antennas for the higher frequency bands. The usual configurations are shown in Figs 3A, B and C, popularly called *quad*, *diamond* and *delta-loop* antennas, respectively. When fed as shown, these antennas are horizontally polarized in the plane orthogonal to that of the loop.

The loops are symmetrical about the center line, shown dashed. If we rotate the loops clockwise through 90°, and if the referenced center line is now the ground plane, we have a *half-loop*, which works together with its image in the ground plane (Figs D, E and F). One end of the half-loop is grounded and the other is fed against ground. The azimuthal pattern at the fundamental resonant frequency is maximum in the plane broadside to the plane of the loop, and the radiated field is vertically polarized. Sometimes wire loops are supported by trees; sometimes by metal masts. Often the mast also supports a Yagi antenna at the top. It has long been known that a metal tower can affect the impedance and radiation pattern of any wire antenna that it also supports. The effect of support structures on the radiation patterns of antennas has been reported, however, for only a few configurations and antenna types.[7]

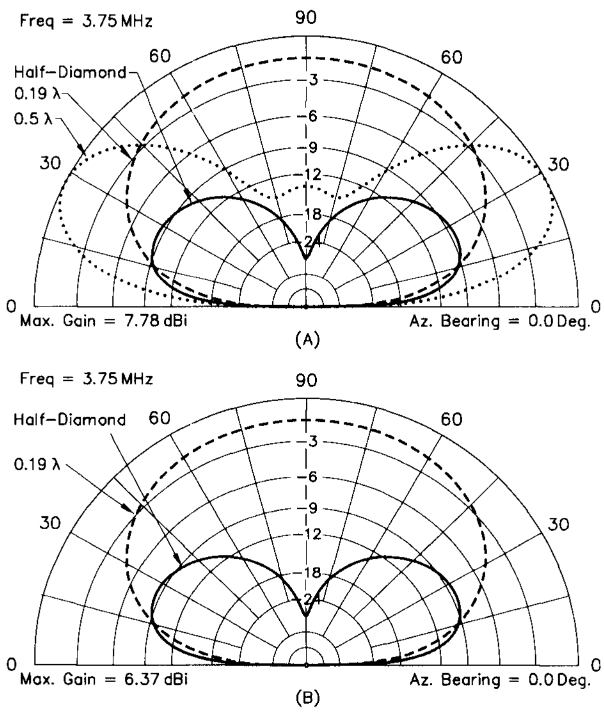

Fig 2—At A, overlay of elevation patterns for 3.75-MHz half-diamond loop, dipoles at 15.24 meters (0.19 λ) and at 40 meters (λ/2) over average ground (σ = 3 mS/m, ε = 13). At B, elevation-pattern overlays for only the half-diamond loop and the dipole at 15.24 meters. See Note 6.

Ufer Ground

Whatever loop configuration is employed, a ground-plane (GP) type loop must be well grounded. My half-delta loop is grounded by a 3-meter ground rod and four buried quarter-wave radials. The fed end is close to the house, so only three radials could have been used. The transceiver is grounded to the power mains ground, which is connected to the copper water-pipe system in the house. Since the house water supply is a well, my water-pipe system is not connected to an extensive and well-grounded city water system. Further, the lead-in pipe from the well is plastic.

One day I decided to tie the backyard chain-link fence into the ground system. This vinyl-coated, steel-fabric fence is fastened to a tubular steel framework, made up of a continuous pipe running the 200-meter perimeter of my backyard. The fence pipe is supported by tubular metal posts set in concrete every 6 meters. A pipe set in concrete in moist earth can make a good ground. Such a ground is known as a *Ufer ground*, after the engineer that studied this method of grounding.[8,9] To ensure a continuous loop of pipe, a buried jumper wire was connected across the gate; and the fence was jumpered to the copper water-pipe system at both sides of the house.

Curiously, the received background noise level (judged

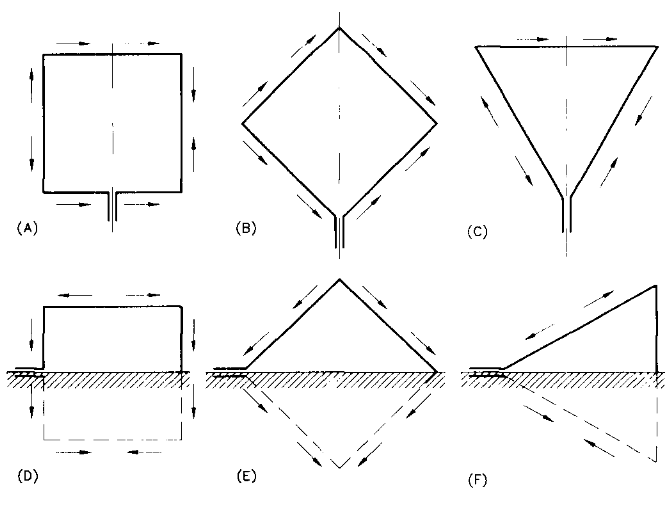

Fig 3—At A, B and C are sketches of horizontally polarized quad, diamond and delta loops. At D, E and F are sketches of vertically polarized ground-plane-type half-quad, half-diamond and half-delta loops.

to be local man-made noise) on the 80-meter band dropped by an S-unit or more when the fence was tied in. I checked this on several occasions. My whole backyard (about half acre) is my ground system!

Initial Experiences

About 15 years ago, after conducting a series of extensive experimental modeling studies, I erected an 80-meter half-delta loop in the back yard.[10] I already had a 15-meter free-standing tower in place (with a 3-meter pipe mast extension to support TV antennas) located just outside the window of the half-basement ham shack. The most logical arrangement was to mount a tree-supported wire loop so that the vertical part of the half-delta loop was as far as possible from the tower, with the sloping wire running toward the tower. But since the wire antenna was in the backyard, the lower end of this sloping wire did not come directly to ground, so that people could walk safely beneath it. This wire ran through an insulator attached to the tower at a height of about 2.5 meters, and from there to ground level. The feed was between the lower end of this wire and ground. See Fig 4A. The spacing between the wire and the tower was about 30 cm.

I noticed that when using the loop there was significantly more of a problem with TVI (recall that the tower also supported TV antennas) than when I used a half-wave drooping dipole suspended from the same tower. Since the half-delta loop is vertically polarized and the tower is about the right height (18.3 meters) to be approximately resonant, I wondered whether this so-called "isolated tower" was in fact isolated, or whether it was a part of the radiating system.

I modeled the antenna and tower, using *EZNEC*. Because the program does not permit any wire to be connected to imperfect ground, I simulated the ground connection for the half-delta loop using two sets of two quarter-wave resonant radials elevated a meter above ground. These radials were directed in the plane orthogonal to the loop. See Fig 4A. For the tower, I simulated the ground connection using four quarter-wave resonant radials elevated 5 mm above ground. [*EZNEC* data files for all these models by VE2CV are located on the Hiram BBS, in an archive file called VE2CV.ZIP—*Ed.*]

According to *EZNEC*, the loop and tower system was resonant at 3.7 MHz (the measured resonant frequency was 3.65 MHz). The tower does carry a significant current, $0.75\angle 69°$ A for a $1\angle 0°$A source current at 3.75 MHz. The current at the grounded far end of the vertical element of the half delta loop is $0.97\angle 69°$ A. Since the currents on the vertical parts of the radiating system (vertical wire and tower) are approximately in quadrature, you might expect a cardioid-like pattern—and this is what *EZNEC* predicts. See Figs 4B and C.

The antenna was oriented so that the plane of the loop was in the N-S plane, with the tower and feed in the north. By chance, my radiation pattern was very suitable for a control station of the Trans Canada Pow Wow Club—with best coverage a bit to the north of the E-W direction. But this was by chance rather than by design, since I had not intended the tower to be a part of the antenna system!

This discovery of this interaction sparked a detailed study, since I wondered whether a half-diamond loop, where the loop is centered on the tower, might be better. First, let's look at some characteristics of half loops in "ideal" situations.

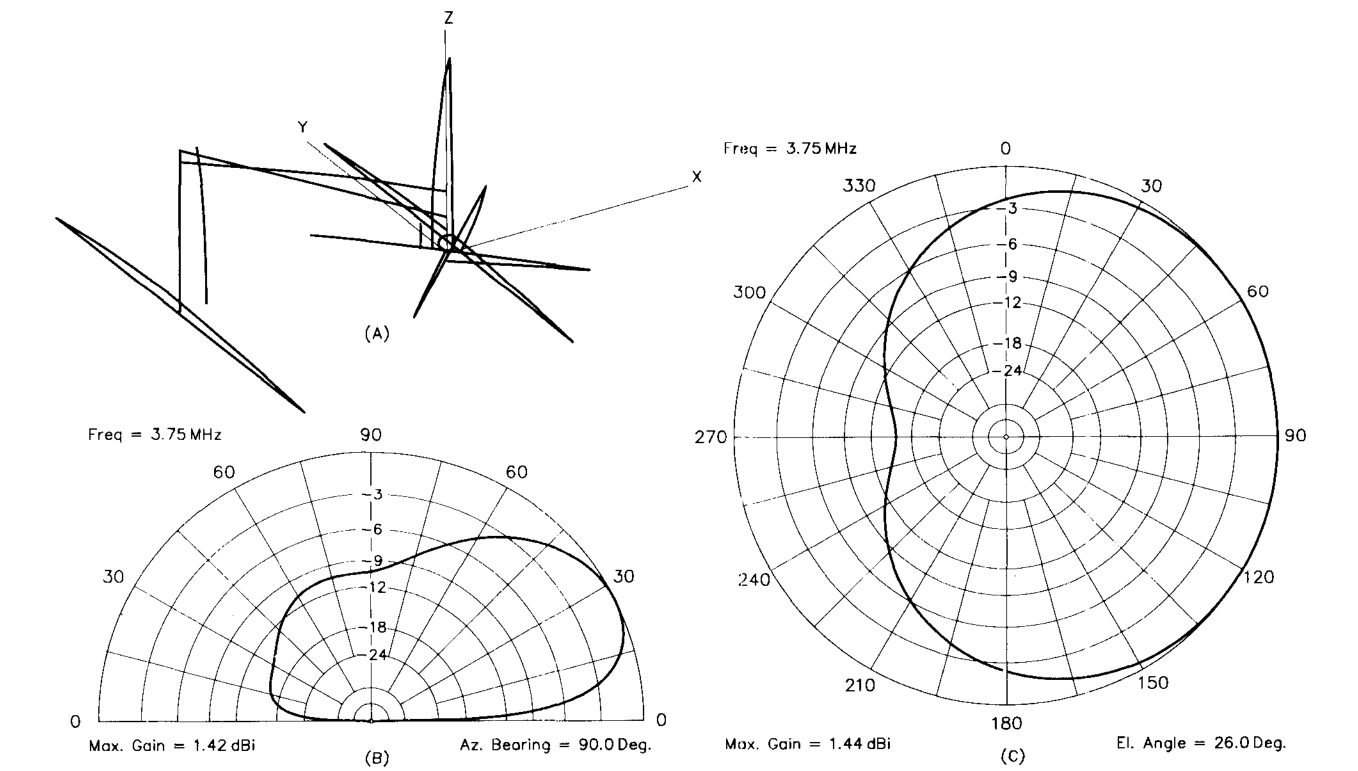

Fig 4—VE2CV's half-delta loop for the 80-meter band. At A, wire model showing currents on the wire loop and the induced current on the tower. At B, elevation pattern and at C, the principal-plane azimuthal pattern. The patterns were calculated for average ground in front of the antenna.

Radiation Characteristics of Half-Loops

Tree Support:

Let's now consider the performance of a tree-supported loop, an "ideal" situation. Again, I wished to predict the gain for a loop over average ground (conductivity σ = 3 mS/m, dielectric constant ε = 13). Again, I used the *EZNEC* program with the loop raised 1-meter off ground, with elevated resonant radials (electrical length $\lambda/4$) to simulate the connection to ground. The fed end of the loop and the grounded end of the loop were connected by a wire. See Fig 5A for the wire model of a half-diamond loop and Fig 5B for a half-quad loop. For an overview on the use of radial wires to simulate ground connection see Note 4.

I modeled half-quad, half-diamond and half-delta loop antennas. There is not a great difference in performance between the various loop configurations. The overhead null is greater for the half-quad and half-diamond loop configuration, compared with the half-delta loop. See Fig 6. The predicted gains are 1.1 dBi, –0.69 dBi and –0.34 dBi, respectively.

The reason why the half quad has the best gain can be seen in Fig 5B by inspecting the current on the antenna. The current on the vertical wires are in phase. Little radiation will result from the small current on the top wire and the base wire carries very little current also. The antenna behaves like a pair of phased verticals.

Tower Support:

When a loop is suspended from a conducting tower, the tower becomes a part of the antenna system. Currents are induced on the tower and the tower reradiates. The far-field radiation pattern is the combined effect due to currents on the loop and the tower. A lattice tower that is 0.237λ or 18.9 meters in height is resonant at a frequency of 3.75 MHz. Clearly, it would be expected to have the maximum induced current, but all conducting towers will have some induced current, since the loop is resonated by cut-and-trim of the loop in the presence of the tower. When you put up an antenna, the usual procedure is to adjust the dimensions of the antenna for resonance in the middle of the band of interest. By so doing, we maximize current on the loop and we also maximize the current on the tower. Similarly, a nonresonant tower will have the smallest effect on the performance of the loop.

Now, let's consider a typical tower installation, where a Yagi is mounted at the top. For the case of a 15-meter tower with a three-element 20-meter Yagi, the self-resonant frequency of the tower plus Yagi is about 2.3 MHz. The tower alone would be self resonant about 4.7 MHz, which is not sufficiently different from the 80-meter band that its presence can be ignored. The presence of a Yagi on top of the tower is beneficial in this case because it actually detunes the system enough to eliminate the interaction.

With no tower present, *EZNEC* tells us that a resonant half-diamond loop with two sets of three resonant elevated radials should have a side length of 0.2627 λ, and the antenna's impedance at resonance is 79 Ω.

Now, if the loop is suspended from a 15-meter metal tower (the minimum tower height needed to support the loop), the loop's impedance changes significantly, from 79 Ω at resonance to $61 + j\,84\ \Omega$. To maintain resonance, the

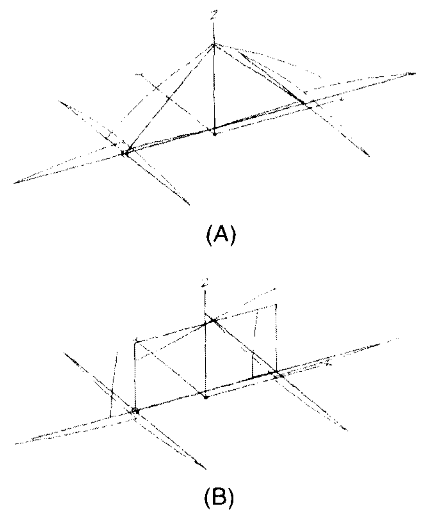

Fig 5—At A, wire model for a 3.75-MHz half-diamond loop, raised off the ground with resonant radials (1 meter high), showing current on the wires. At B, same for a half-quad loop. For clarity, currents are plotted as amplitude only, without regard to phase.

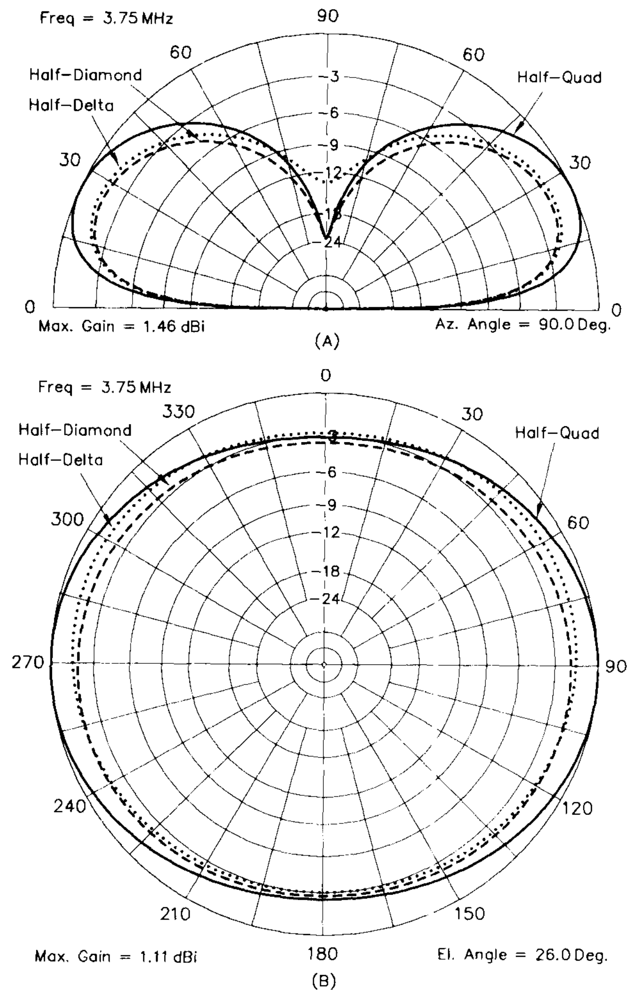

Fig 6—Radiation-pattern comparisons for a half-quad (solid line: gain 1.2 dBi), half diamond (dashed line: gain –0.2 dBi) and half delta loop (dotted line: gain –0.03 dBi) over average ground.

size of the loop must be decreased. The tower's base current is significant, at $0.86\angle 178^\circ$ A (for a source current on the loop of $1\angle 0^\circ$ A). Fig 7A shows the current on the loop antenna and the 15-meter high tower.

The tower and the loop are closely coupled, since if we resonate the loop by reducing its size, the current on the tower is also maximized. We can even interchange the source from the loop to the tower. Indeed, the metal tower is part of the antenna system.

For a resonant tower, with a height of 18.9 meters at 3.75 MHz, the tower is very closely coupled with the loop. The loop's impedance is drastically altered, to $217.3 + j\,81.3\ \Omega$, and the current at the tower base is large, at $1.81\angle 98.9^\circ$ A.

How does this affect the radiation patterns? Fig 8 shows superimposed patterns for the loop by itself, for the loop with a 15-meter tower and for the loop with an 18.9-meter tower. Since the resonant tower carries the most current, the pattern is omnidirectional—in fact, *you might just as well dispense with the loop and feed the tower*!

The closely coupled loop-tower system complicates tuning, impedance and pattern prediction. For a nonresonant loop that is tuned with an antenna tuner, the azimuthal pattern can be almost anything, depending on loop dimensions, frequency, and the height and top loading of the tower. It can even have a null in an unexpected direction.

For a typical 15-meter tower with a 20-meter Yagi beam on it (see Fig 7B), the loop's resonant frequency is close to that for a tree-supported loop. The impedance rises to $111 + j\,16\ \Omega$. Because of the top loading of the Yagi, the resonant frequency for the system is the same as for a tree-supported antenna, with an impedance of 82 Ω. While the tower does carry a small current at $0.34\angle 51^\circ$ A, the pattern and gain are not significantly changed.

A half-delta loop is often connected to a tower, using it as part of the antenna. If this tower also supports a 20-meter Yagi, the loop antenna's impedance and radiation pattern can be changed dramatically. A half-delta loop dimensioned for resonance at 3.75 MHz on a 15.24-meter mast without a Yagi has, according to *MININEC*, an impedance of 58 Ω. If the tower supports a typical three-element 20-meter Yagi, the impedance of this loop is changed to $67 - j\,524\ \Omega$. This configuration exhibits resonances at 2.91 and 2.45 MHz, which agree with the 3.0 and 2.44-MHz resonant frequencies scaled from our earlier experimental scale model measurements. (For this analysis I used *MININEC* since the experimental model measurements were made over a metal ground plane. See Note 10.) The system can be resonated by lengthening the sloping wire to 44.7 meters, but the radiation pattern bears little resemblance to that of a half-loop antenna. The pattern is more like that for a half-sloper antenna.

Similar problems are found if we reverse the fed and ground ends of the half-delta loop and isolate the loop from the tower and Yagi, an arrangement suggested by one of my correspondents. The end of the sloping wire is now remote from the tower and connected to ground, and the vertical part of the wire loop is close to the tower, but insulated from the tower. The feed is between the lower end of this wire and ground. With this arrangement we are

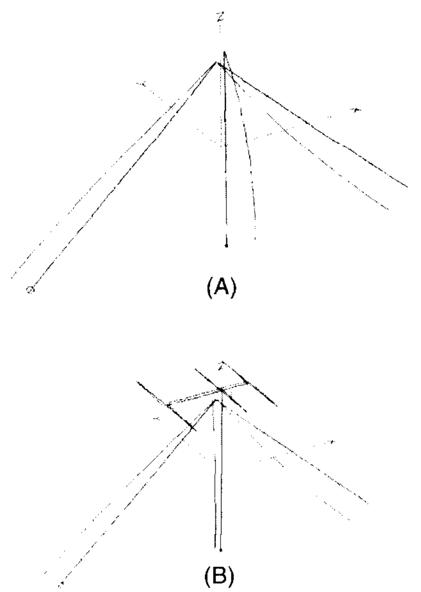

Fig 7—Wire model and current distribution for a resonant half-diamond loop. At A, the loop is suspended from a 19.8-meter tower. At B, the loop is suspended from a 15-meter tower with a 3-element 20-meter Yagi.

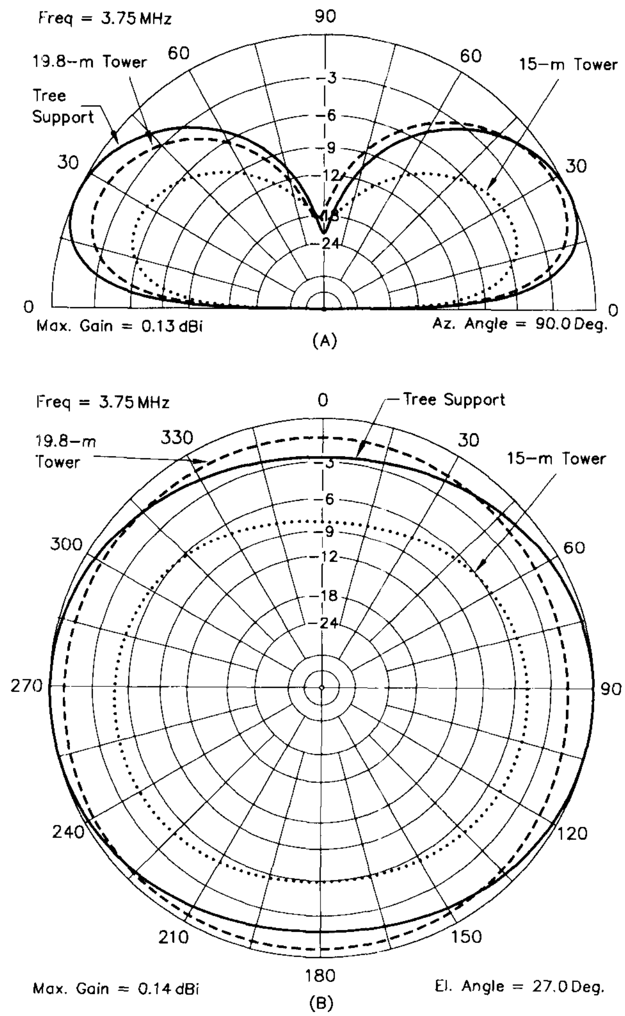

Fig 8—Radiation-pattern comparisons for a 3.75-MHz half-diamond loop with tree support (solid line), on a 15-meter tower (dashed line) and on a 19.8-meter tower (dotted line).

ensuring close coupling between the loop and the tower! The current induced on the tower is out of phase with the current on the parallel vertical wire forming the half-delta loop—this is expected and is a transmission-line effect. The antenna's impedance and pattern are significantly changed. It is not possible with the half-delta loop to escape a dominant influence of a metal supporting tower on the pattern, since the loop is either directly connected to or closely coupled to the tower.

Clearly, if a metal tower is used to support a half loop, the half-diamond loop is the preferred arrangement, providing that the tower is not resonant in the band of interest.

How Many Radials Do You Need?:

To model GP-type half-loop antennas using *EZNEC*, I also used elevated resonant radials. Again see Fig 5A, which shows a half-diamond loop with two sets of three elevated resonant radials. These are 0.2333-λ long for a loop 1 meter over average ground. I also added a wire connecting the lower ends of the sloping sides. The side length for a half-diamond loop is an electrical λ/4 wavelength, or 0.2627 λ for this loop. The current is maximum at the fed and ground ends of the loop. If these currents are exactly equal and in phase, there should be little or no current on the wire connecting the lower ends of the loop. The phase difference between the current on one end and on the other end of this wire should be 180°. According to *EZNEC*, the current on each end of this loop wire is indeed small, at 0.12 A (0.02 A for a half-quad with elevated radials), compared to the source current of 1 A. The phase difference between current on the ends of this wire is 179°.

The predicted gain for this antenna over average ground at 3.75 MHz is –0.7 dBi, and the resonant impedance is 85 Ω. The resonant radials do indeed simulate an effective ground connection since the loop behaves like a GP-type resonant half loop. Fig 5A also shows the current distribution on the wires. Do not expect current on the wire connecting the ground ends of the loop to contribute much to the radiated field. They could be eliminated; see Fig 9A. There is only a small change in impedance (to 79 – *j* 2 Ω) and a small change in gain (to –0.43 dBi). While the antenna may no longer look like a half loop, it has the characteristics of a GP loop.

Since the antenna radiates predominantly in the broadside directions, perhaps radials are needed only in the broadside directions. See Fig 9B. Again, there is a change in impedance (to 71 – *j* 13 Ω) and a change in gain (to a maximum of 0.38 dBi). Fig 10 shows the radiation patterns for the antennas in Figs 5A, 9A and 9B.

Finally, perhaps the radials on one of the broadside directions can be eliminated to create a directive antenna system. See Fig 9C. The antenna impedance is changed (to 110 + *j* 15 Ω) and the gain is changed (to 0.8 dBi at a launch angle of 30° at the azimuth of maximum gain). Fig 11 shows the radiation pattern for the antenna shown in Fig 9C. Notice that this system is unidirectional.

One of my colleagues, Bob Eldridge, VE7BS, erected a half-diamond loop for 160 meters. Since his ground is rocky he decided to raise the loop, cut the base wire in the middle and swing the two ends of the wire at right angles to the plane of the loop. See Fig 12. The apex is at 33.5 meters and

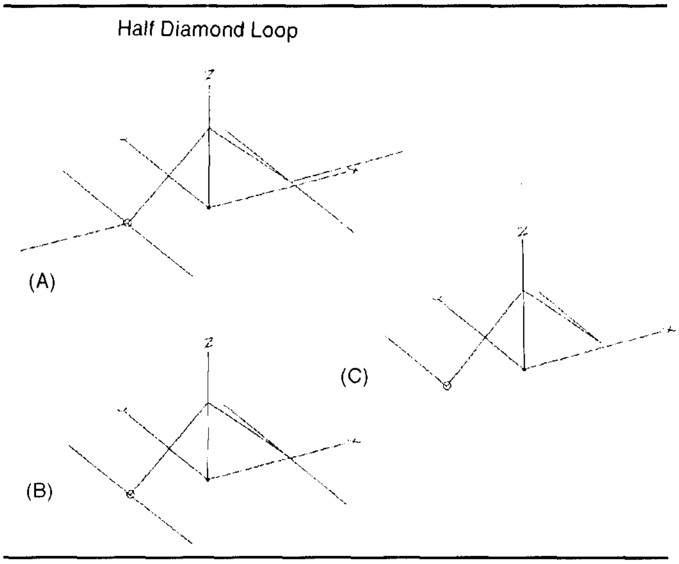

Fig 9—Wire model for a 3.75-MHz GP-type half-diamond loop elevated with resonant radials. At A, without wire connecting the lower ends of the sloping sides. At B, without radial wires in the plane of the antenna. At C, with radial wires on only one broadside direction.

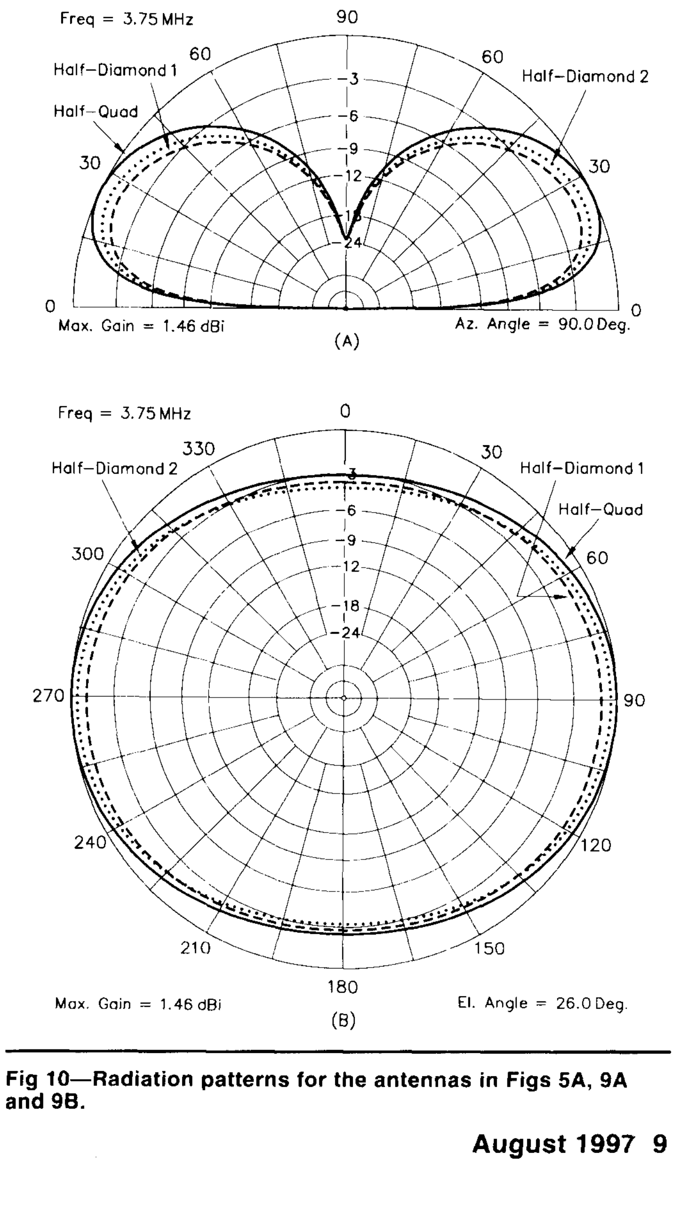

Fig 10—Radiation patterns for the antennas in Figs 5A, 9A and 9B.

the wires sloping down from the apex are each 40.8 meters long. The ends of the sloping wires are 4.6 meters above ground. He feeds the antenna with coax and a choke balun. The center conductor of the coax is connected to the sloping wire, the shield to the 39.6-meter wires running at right angles to the loop (which are now radial wires, 4.6 meters high). The radial wires are also sloped, with an end height of 1.5 meters. The antenna was resonant at 1.84 MHz, with an SWR about 2:1. According to *EZNEC*, the resonant frequency should be 1.86 MHz, with an impedance of 105 Ω, a gain of 1.4 dBi, a take-off angle of 27° and a front-to-back ratio of 4.2 dB.

Lifting the half-loop off the ground apparently made a quieter receive antenna. The resonance was easier to establish and to maintain and the gain definitely favored the direction toward which the radials run. VE7BS noted that amateur stations in the favored direction, Japan, were always better received than on a horizontal loop, but stations in the opposite direction (southern and southeastern USA) were usually worse.

Good Ground Conductivity is important—But to What Distance?

The ground conductivity at distances a long way in front of the antenna is important for low launch angles if the polarization is vertical—to about 50 λ. This is about 4 km for the 80-meter band. On the map in Fig 13 a 50-λ radius (4 km) circle is drawn around my QTH. Inside this circle, in the west through northwest and northeast through east directions, the ground conductivity is estimated to be reasonably good (5 mS/m). Toward the west to distances beyond 50 λ there is forest land, suburban countryside and farms, and the Ottawa River. Toward the east are two golf courses. Just beyond this distance, however, the easterly sector radial path encounters the cities of Hull/Ottawa. Cities are characterized by a low effective ground conductivity.

During the past five years I've made countless tests of signal strengths received on my grounded half-delta loop compared with those on a 15-meter-high dipole. Toward the west for distances beyond about 1000 km, the signal strengths received on the half-delta loop are usually stronger, by as much as 1 to 2 S units for west coast signals. As you'd expect, there are variations due to different propagation conditions and sometimes the difference is not so marked. But the signals on the half-delta loop are never less than those received on the dipole, and the signal-to-noise ratio is generally improved.

Toward the east the opposite is found. The signal strengths for stations in Newfoundland and the UK are always stronger on the dipole. This must validate the principle that the ground beyond 50 λ is important for the launch of low-angle sky-waves. The low conductivity of the city environment beyond this distance must be influencing the ability of the antenna to receive low-angle skywaves. I suspect that this must be the explanation for the apparent "unidirectional" characteristic of the my half loop.

I've also found the same EW-WE asymmetry using another vertically polarized antenna, my compact 1.7-meter diameter loop (an AMA 11). On many nights I've found reception to be poor for west coast stations on 3750 kHz, because of noise and interference, using either the dipole or the 80-meter half-delta loop. On switching to the 1.7-meter loop reception was clear—not just marginally so, but significantly better. The received signal strengths were reduced by an S unit or two when using the small loop, but the noise and interference was reduced even more, improving the S/N. Switching from the dipole to the half-delta loop and then to the compact loop improved reception progressively—but I've never seen this improvement for stations coming from the east. When propagation conditions are

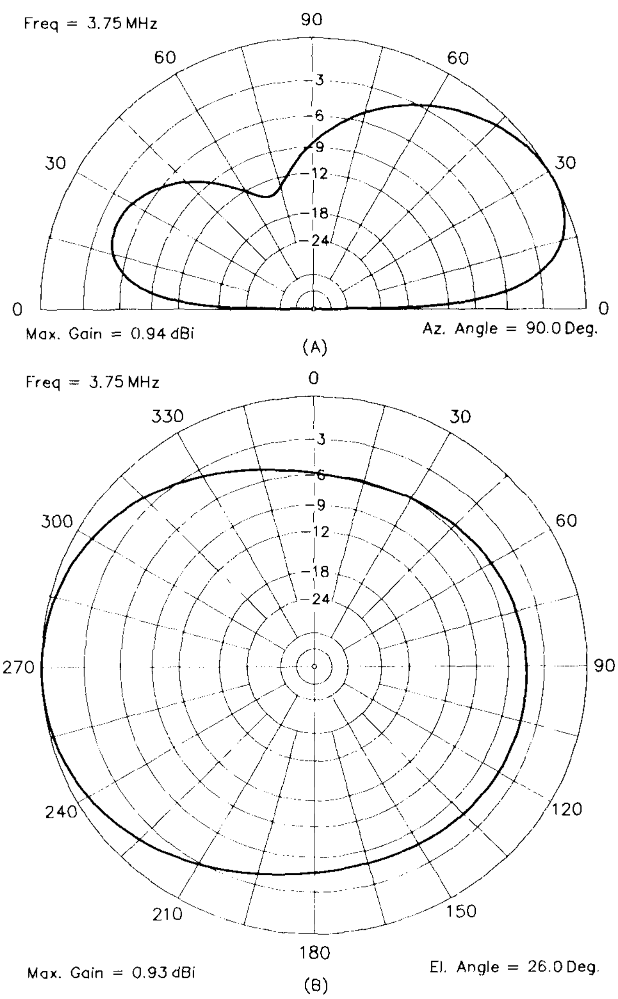

Fig 11—Radiation patterns for the antenna in Fig 9C.

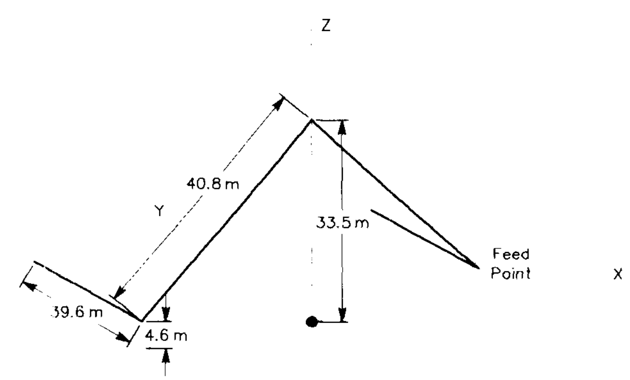

Fig 12—Model of VE7BS's modified half-diamond loop for 160 meters.

good, stations in NW Ontario have able to hear me using the little 1.7-meter loop, even as far west as the west coast (Salt Spring Island). On the other hand, I never received good reception reports from stations in Nova Scotia using either loop.

I've been lauding the advantages of using a GP-type loop for a decade and a half—not as an end-all, be-all antenna, but as a complement for a dipole for 80 and 160 meters. Elevating the antenna system above head height (2.5 meters), results in a quieter receiving antenna and also makes the installation of many on-the-ground or buried radials unnecessary. Further, the use of resonant radials provides some control over the antenna's directivity.

The fact that a tuned loop can be so closely coupled to its supporting tower is a surprise to many. In retrospect, this should have been anticipated. This completes my overview of vertically polarized half-loop antennas. Now let us consider full-wave delta loops.

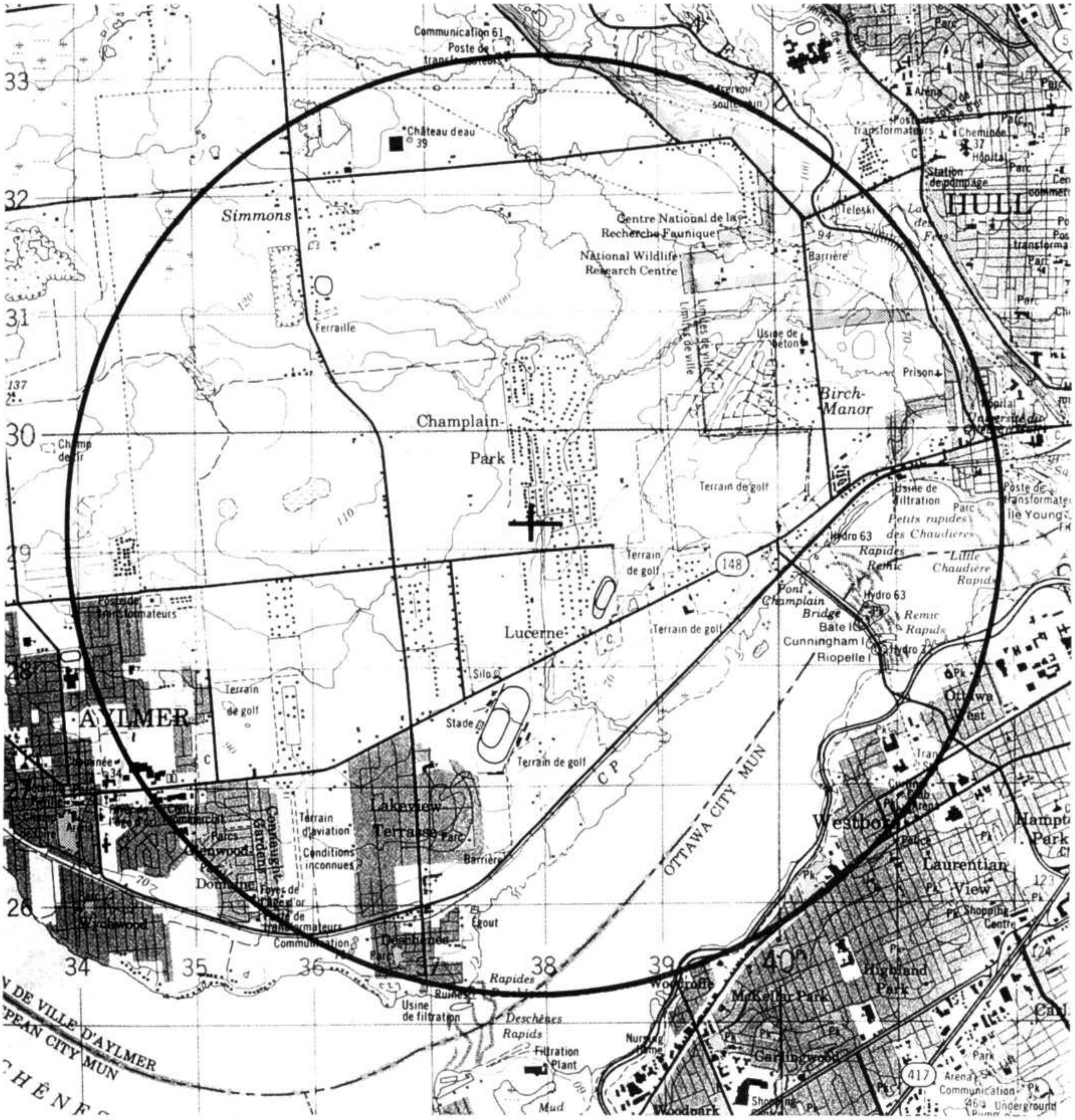

Fig 13—Map showing VE2CV's QTH, with a circle drawn at 4 km. This corresponds to a 50-λ radius circle at 3.75 MHz.

Full-Wave Delta Loops

Earlier, I discussed how full-sized loops can be delta, diamond or quad in shape. Here we will consider the full-wave delta loop configuration in more detail, since this form can easily be suspended from a single pole or tower.

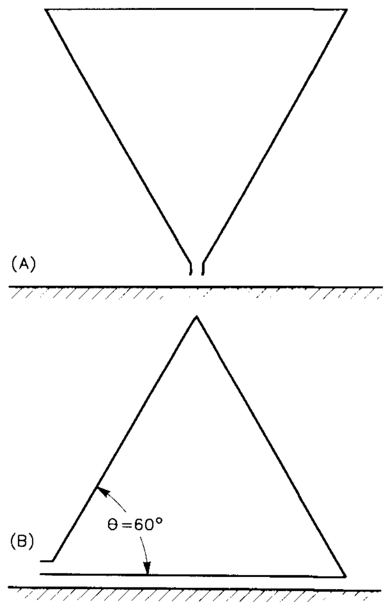

Fig 14—At A, a conventional apex-down, apex-fed delta loop for horizontal polarization. At B, a basic delta loop configuration for DXing, with an apex-up, lower-corner feed.

A wire loop that has attracted considerable interest for low-band DX is the apex-up lower corner-feed delta loop.[11,12] This is a loop that has been rotated through 60° with respect to its orientation for horizontal polarization (see Figs 14A and 14B), rather than 90° as is the case for ground-plane type half-loops. Therefore, the radiated field should be a mixture of both horizontal and vertical polarization components.

This could be considered a disadvantage, since to hear distant stations it is advantageous to have an antenna system that has a pattern null overhead to reduce the strength of strong local signals, man-made radio noise and interference. Another important practical disadvantage is that a high tower is required—about 27 meters (88.5 feet) is needed for 80 meters. This height assumes that the bottom horizontal wire is 2.5-meters high, so you can walk beneath the wire.

More typically, you might have a tower that is shorter than 27 meters, so the delta loop must be 'squashed' to fit on a shorter mast. This makes the base longer than the sloping sides. Squashing the loop lowers the resonant impedance of the loop and changes its radiation pattern.

The loop is full-wave resonant, so no ground connection is needed. In fact, the loop should be completely isolated from ground employing a current balun for feed using a coaxial cable transmission line. I stress that you should use a *current* balun because the loop is not symmetrical with respect to ground. The impedance and radiation patterns in this article were computed using *EZNEC* using the Sommerfeld-Norton ground model. Proper account is taken of the ground beneath the antenna, as well as in front of the antenna, and a current source is used, isolated from ground.

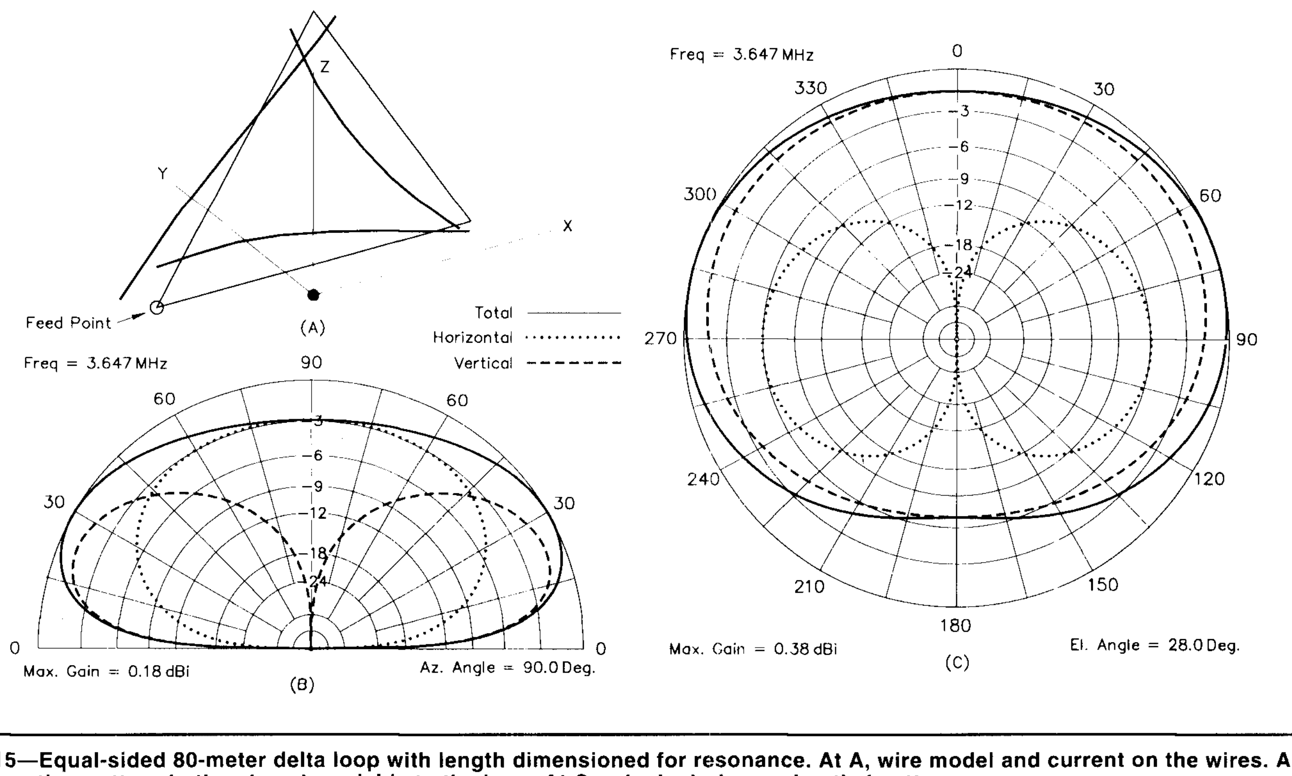

Fig 15—Equal-sided 80-meter delta loop with length dimensioned for resonance. At A, wire model and current on the wires. At B, elevation pattern in the plane broadside to the loop. At C, principal-plane azimuthal pattern.

Effect of Loop Shape on Radiation Characteristics

Tree or Wooden Pole Support:

If the shape of the delta loop is an equilateral triangle—that is, all sides are equal, $\theta = 60°$, where θ is the angle between a sloping side and the horizontal base of the triangle—the loop should have a radiation field that is a mixture of both horizontal and vertical polarization. In Fig 15 you can see that the antenna is dominantly vertically polarized. The azimuthal pattern is not symmetrical in the plane of the loop, with maximum gain in the direction away from the feed. The antenna's resonant impedance (for a height 2.5 meters over average ground) is 164 Ω.

As the loop is squashed, its resonant impedance decreases, the principal-plane azimuthal pattern becomes more symmetrical and the vertically polarized component more dominant. The resonant impedances of delta loops where θ equals 31°, 35°, 40° and 60° are 48, 62, 78 and (as noted above) 164 Ω, respectively. The resonant impedance, and dimensions for resonance, depend on the height of the loop and on the ground conductivity.

The radiation characteristics for the 35° loop will be described in detail, since this delta loop configuration can be supported by a 15-meter (50-foot) wooden pole or tree. See Fig 16. According to *EZNEC*, the dimensions of this loop are: length of sloping sides 23.073 meters; length of base wire 37.8 meters; perimeter 1.05 λ. Notice that the horizontally polarized component of the radiation field is negligible. The reason for this can be seen by examining the currents on the wire structure in Fig 16A. The currents each side of center on the horizontal wire, or the base of the delta loop, are oppositely phased. Hence radiation from this wire is negligible. This results from the fact that the length of the wire is almost a halfwave (0.47 λ). The sloping arms (length 0.29 λ) carry approximately equal in-phase currents. The antenna's radiation characteristics are similar to a pair of phased vertical monopoles.

Tower Support:

As noted earlier, if the supporting mast is a metal tower, current will be induced to flow on the tower, and this significantly changes the impedance of the loop and its pattern. The effect of the tower depends on loop shape, tower height and whether or not the tower is well grounded.

Let's consider a case study where the effect of the tower is significant; that is, a 35° delta loop suspended from a resonant tower. To simulate a realistic, well-grounded tower using *EZNEC*, I used a resonant vertical wire with eight close-to-the-ground resonant radials, 5 mm off the ground surface. The physical height of the wire simulating a metal tower is the height for which the electrical length of the wire and the tower are the same.

With the resonant tower, about 19.8-meters high, the loop's resonant frequency is reduced from 3.75 MHz to 3.59 MHz. Figs 17B and 17C show the radiation patterns for a resonant loop. The loop size had to be reduced by 4.4% for resonance at 3.75 MHz. As noted previously for the half diamond loop (Fig 4a), the tower caries a significant current, 1.7∠137° A. In fact, the current on the tower dominates—the azimuthal pattern is essentially omnidirectional. Adding more radials to simulate a better ground for the tower causes the tower current to increase, and the azimuthal pattern becomes even more circular. For comparison, Figs 17B and 17C show the patterns superimposed

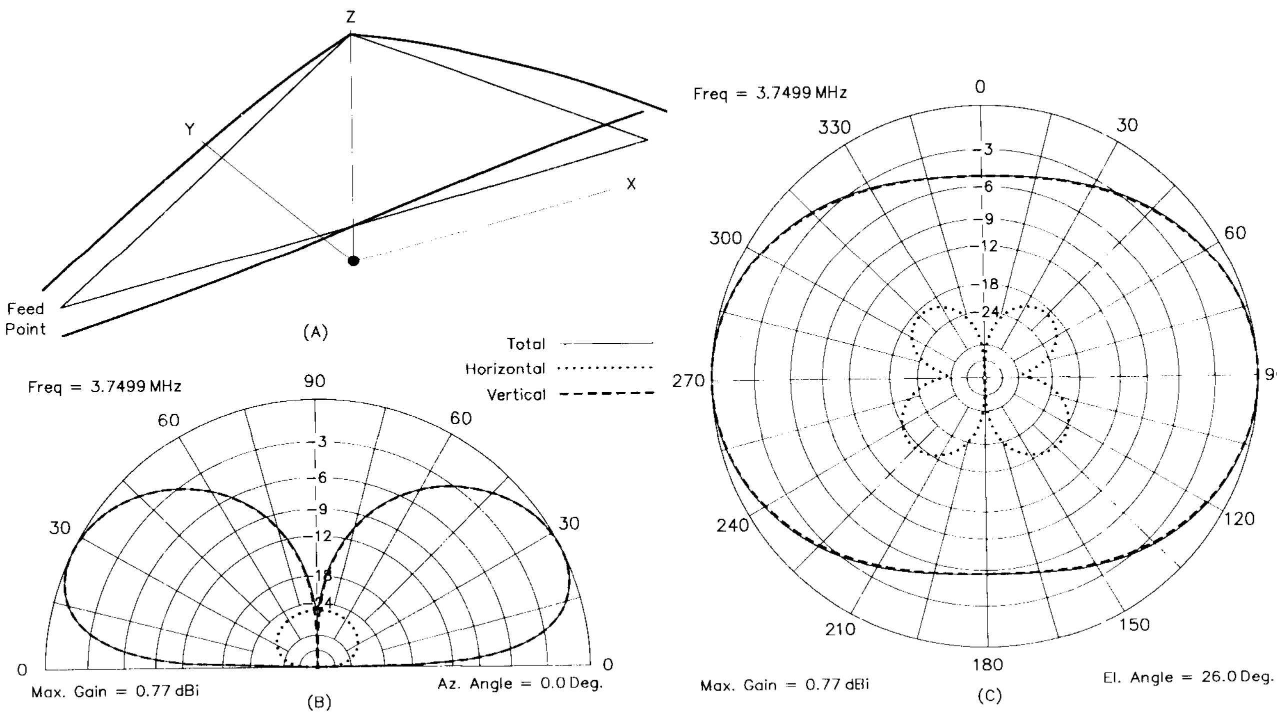

Fig 16—Delta loop squashed so that it can be suspended from a 15-meter wooden pole or tree. Length of sloping sides is 23.072 meters; base length is 37.8 meters, $\theta = 35°$. Resonant frequency, according to *EZNEC*, is 3.75 MHz. At A, the wire model and current on the wires are shown. At B, the elevation pattern in the plane broadside to the loop is shown. At C, the principal-plane azimuthal pattern is shown.

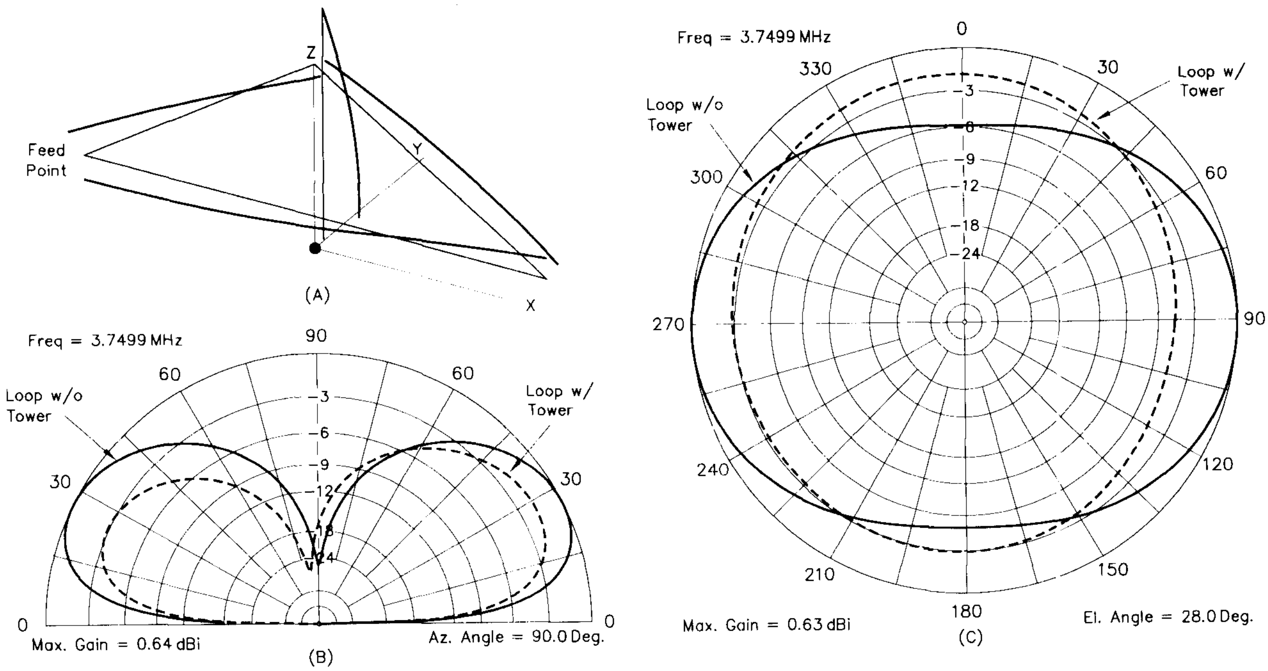

Fig 17—Patterns for antenna in Fig 16, compared with the patterns for a loop supported by a resonant metal tower. Patterns plotted are the principal-plane patterns for resonant dimensions. Note that in this case study current induced to flow on the conducting tower dominates, since the azimuthal pattern is more or less omnidirectional.

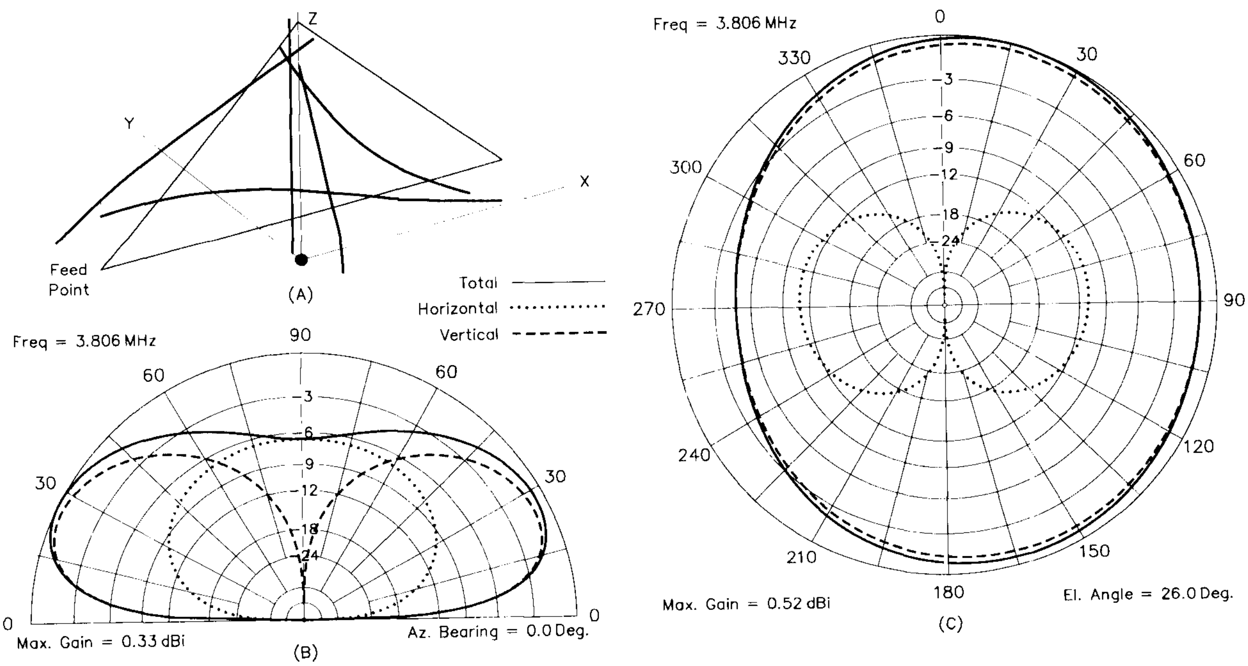

Fig 18—At A, wire model and wire currents for ON4UN's delta loop at resonant frequency of 3.88 MHz. At B, elevation pattern in the plane broadside to the loop is shown. At C, principal-plane azimuthal pattern is shown.

for the case where the tower is nonconducting (a wooden pole or tree support).

John Devoldere, ON4UN, in his book on 80-meter DXing, describes an 80-meter apex-up lower-corner-feed delta loop that has undoubtedly been duplicated by many. He sketched a loop with the following dimensions: sloping sides 23.75 meters, base length 33 meters, $\theta = 46°$. The loop was shown suspended from a 21-meter (lattice) tower and is shown fed directly with 50-Ω coaxial cable. While no balun is employed, his sketch suggests buried coaxial cable. The shield of the coax is connected to the horizontal wire, the inner conductor to the sloping wire. The SWR he shows for this antenna, which is apparently resonant at 3.8 MHz, is very low. He also reports that the delta loop has little directivity at high elevation angles, but shows very pronounced nulls at low elevation angles (–30 dB). The loops I have modeled for $\theta < 46°$ have little gain at high elevation angles and the patterns do not show such pronounced nulls on the ends.

The loop shown in Fig 18A is modeled after ON4UN's loop. For a nonconducting support (tree or wooden pole), the resonant frequency for the loop dimensioned as above (according to *EZNEC*) is 3.93 MHz and the impedance at resonance 96 Ω (for the case of a loop over average ground). With a 21-meter metal tower and perfectly conducting ground, an *EZNEC* analysis shows the predicted resonant frequency at 3.8 MHz and the impedance at 128 Ω. For average ground, a conducting tower with eight on-the-ground resonant radials (radial height 5 mm) the computed resonance is at 3.88 MHz and the impedance 145 Ω. The base current on the tower is indeed high ($0.67\angle{-118°}$ A) and the gain 0.78 dBi.

It seems curious that users of this antenna consistently use direct coax feed, and apparently manage to achieve a low VSWR, since, whatever configuration is modeled, wooden pole or conducting tower, the predicted VSWR for direct 50-Ω feed would be 2 or 3:1.

Effect of Loop Size:

The loop was resonant for the patterns so far described. However, if a loop is dimensioned for the middle of the 80-meter band, say for 3.75 MHz, it could also be used with an antenna tuner for operation on the upper and lower ends of the band. Fig 19 shows the patterns at 3.5 MHz for the 3.75-MHz loop shown in Fig 17, $\theta = 35°$. Notice the change in pattern. The horizontally polarized component, which is almost completely canceled when the loop is operated on its resonant frequency, is not canceled to the same degree when the loop is operated at a frequency significantly below its resonant frequency. The current in the wires is no longer symmetrical. The 4-MHz pattern (not shown) is less affected.

Concluding Remarks on Delta Loops:

The impedance and pattern of full-wave delta loops depends markedly on the shape of the loop, whether it is operated on its resonant frequency and on whether it is supported by a wooden pole or metal mast. The impedance and gain depend on the height of the loop and the ground conductivity. A tower height that is approximately resonant in the frequency band of interest should be avoided if possible.

These characteristics undoubtedly explain observations that have puzzled me in the past. Sometimes I've noticed that when a distant amateur station using a delta loop has switched from a dipole to the loop the result can be quite different. In some cases, the near vertical incidence sky-wave signal decreases by several S-units; in other cases the change is not so dramatic. The reverse can be observed in the case of distant DX signals; that is, the loop can be better than the dipole, depending on the height of the dipole. Clearly, the dimensions, shape and size of the loop and whether the loop is operated on its resonant frequency can explain these observations.

Since the loop is isolated from ground, and is not sym-

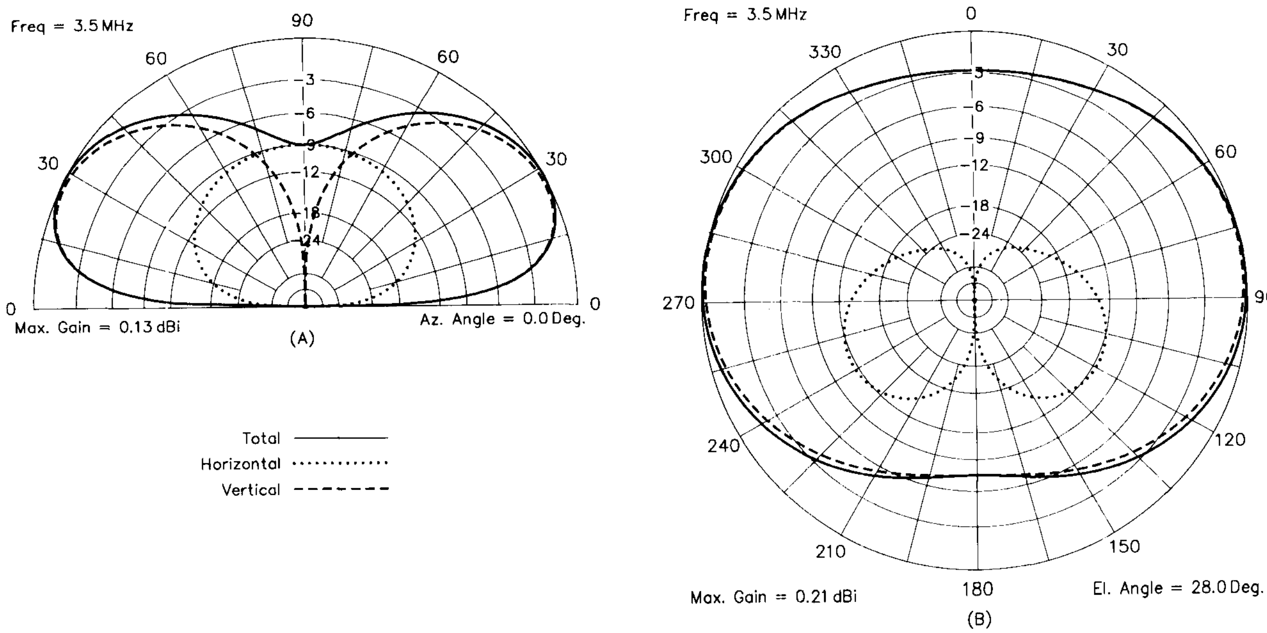

Fig 19—Patterns for antenna shown in Fig 17, for a frequency below resonance, 3.5 MHz.

metrical with respect to ground, it should be fed by a current balun to force the current to be distributed properly.

Notes

[1]Belrose, J. S., "The Half-Delta Loop," *HR Magazine*, May 1982, pp 37-39.

[2]Belrose, J. S., "An Update on Compact Transmitting Loops", *QST*, Nov 1993, pp 37-40.

[3]Christman, A., "Elevated Vertical Antenna Systems," *QST*, Aug 1988, pp 35-42.

[4]Belrose, J. S., "Ground-Plane Type Antennas with Elevated Radial Systems", *Proceedings of the Twelfth Review of Progress in Applied Computational Electromagnetics*, Naval Post Graduate School, Monterey, CA, 18-22 Mar 1996, pp 2-11.

[5]*EZNEC Antenna Analysis Program*. Available from Roy Lewallen, W7EL, PO Box 6658, Beaverton, OR 97007.

[6] Note: The complex permittivity that Burke, Miller and others (developers of the NEC code) use is: real part = relative dielectric constant; imaginary part = 60 × wavelength in meters × conductivity in S/m. Thus the complex permittivity for average ground ($\sigma = 3$ mS/m, $\varepsilon = 13$) at a frequency of 3.75 MHz is $13 - j\,14.4\ \Omega$. Burke has shown [in private communication, 1995] for monopole antennas with radial wire systems, that ground induced losses are a maximum when the loss tangent is one; that is, the real and imaginary parts of the complex permittivity are equal. This is close to the case for 3.75 MHz, and so average ground may not result in average ground-induced loss!

[7]Belrose, J. S., "The Effect of Supporting Structures on Simple Wire Antennas," *QST*, Dec 1982, pp 32-35; Feedback, *QST*, Jul 1983, p 41. Addendum, *QST* (Technical Correspondence), Oct 1984, p 42.

[8]Ufer, H. G., "Investigation and Testing of Footing Type Grounding Electrodes for Electrical Installations," *IEEE Trans. on Power Apparatus and Systems*, No. 83, Oct 1964, pp 1042-1048.

[9]Fagan, E. J., and Lee, R. H., "The Use of Concrete-Enclosed Reinforcing Rods as Grounding Electrodes," *IEEE Trans. on Industry and General Applications*, No. IGA-6, Jul/Aug 1970, pp 337-348.

[10]Belrose, J. S., and DeMaw, D., "The Half-Delta Loop: A Critical Analysis and Practical Deployment," *QST*, Sep 1982, pp 28-32.

[11]Mayhead, L., "Loop Aerials Close to the Ground," *Radio Communications*, May 1974, pp 298-301.

[12]Devoldere, J., *80-M DX Handbook*, Communications Technologies, Inc, Greenville, NH, 1977. A revised (1994) Second Edition is available, see current ARRL Publications Catalogue.

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Synthesizing Vacuum Tubes

Put that old tube-type receiver back to work by replacing the burnt-out tubes with JFET circuits.

By Parker R. Cope, W2GOM/7

Some wag once said that a vacuum tube was just an N-channel depletion-mode FET with a light in it to tell you when it was good. There is enough truth in the comparison to warrant exploring the use of FETs in hard-to-fill receiver tube sockets. Of course, a one-to-one replacement isn't possible, but with a little headscratching, an FET-based substitute may be built for either a pentode or triode vacuum tube. The substitute costs less and has a longer life expectancy than a tube, and there's no penalty in performance.

Comparing Tubes and JFETs

A pentode vacuum tube has five elements: cathode, control grid (G1), screen grid (G2), suppressor grid (G3) and plate. The plate current is essentially independent of plate voltage; the dynamic plate resistance is very high. This high plate resistance does not load a resonant output tank, so simple tuned output circuits can be used. And the low capacitance between G1 and the plate results in little coupling from output to input, so no special circuitry is needed for stable amplification at RF.

A triode has only three elements: a cathode, a control grid and a plate. The plate current is a function of plate voltage and grid voltage; the dynamic resistance is relatively low. The coupling between output and input through the plate-to-grid capacitance is relatively tight, which usually restricts triodes to use in untuned applications. They can be used at RF with special circuits or under special conditions, but are more often found in audio or other low-frequency applications.

A receiving tube typically requires plate voltages of 100 V or so, positive with respect to the cathode. The grid is operated negative with respect to the cathode, and plate current increases as the grid-to-cathode voltage becomes less negative. Grid current is typically a few microamps when the grid is negative but can be several milliamps when the grid is driven positive with respect to the cathode. Plate currents in receiving tubes are a few milliamps, and output power is normally not a significant consideration.

The major parameters of tubes that we'll be concerned with are μ, r_p and g_m, which are related by:

$$\mu = g_m r_p \qquad \text{Eq 1}$$

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where g_m (transconductance) is the change in plate current for a change in grid voltage, with plate voltage held constant; r_p (plate resistance) is the change in plate current for a change in plate voltage, with grid voltage held constant; and μ is the change in plate voltage needed to hold plate current constant for a change in grid voltage—in a word, gain. Pentodes usually specify g_m and r_p, while triodes specify μ and r_p. g_m is typically in the range of 1000 to 10,000 μmhos. [The more modern term for the mho is the *siemens*. Receiving tube databooks, however, are likely to give g_m in μmhos rather than in μS.—*Ed.*] r_p is in the range of 1 MΩ for pentodes and less than 100 kΩ for triodes.

The N-channel depletion mode JFET has features in common with vacuum tubes. The drain (equivalent to the plate) operates with a voltage that's positive with respect to the source (equivalent to the cathode). Drain current is essentially independent of drain voltage; the dynamic drain resistance is very high. The gate (equivalent to the control grid, G1) operates at a voltage that's negative with respect to the source. Gate current is a nanoamp or so (essentially zero) when the gate is negative with respect to the source, and maximum drain current occurs with 0 V gate-to-source. The gate current can be several milliamps when the gate is positive with respect to the source. Of course, the JFET operates with much lower voltages than tubes, from a few volts to 25 or 50 V maximum.

The JFET has only a single gate, and the gate-to-drain capacitance is relatively large, like the triode grid-to-plate capacitance, which limits its RF applications to special circuits or operation under special conditions. There are dual-gate MOSFETS (eg, 3N200, 3N201, 3N140, 3N187) that have low output-to-input coupling and are used in TV front ends without special circuits. While their RF characteristics may match a pentode, their transconductances may not. A JFET can have a transconductance much greater than any tube, and its operating point can be adjusted to produce various values of g_m.

Two discrete JFETs or tube triodes connected in cascode, as shown in Fig 1, have excellent RF properties. In the JFET cascode, Q1 is a common-source amplifier that drives the source of the common-gate amplifier Q2. The voltage gain of Q1 is $g_{fs(1)}R_s$, where $g_{fs(1)}$ is the common-source transconductance of Q1, and R_s is the source impedance of Q2, which is approximately $1/g_{fs(2)}$. When the g_{fs} values of both transistors are equal, the voltage gain of Q1 is g_{fs}/g_{fs}, and Q1 is unconditionally stable. Since the current in Q2 is the same as the current in Q1, the gain of Q2 is $g_{fs}R_L$, where R_L is the impedance in the drain of Q2 (the load). The cascode amplifier is stable, even though the gain of Q2 (and thus the overall gain) may be high, because the grounded gate of Q2 effectively shunts its C_{gd} to ground. The capacitance between output and input is generally in the range of femtofarads.

Mimicking the Tube's Parameters with JFETs

As you can see the basic nature of a pentode or triode vacuum tube can be matched using JFETs. The next step is to make our JFET circuit exhibit the same parameters—from Eq 1—as the vacuum tube we're trying to replace. What we'll do is trim the transconductance of a JFET to the particular value needed by adjusting the JFET's operating point.

To simulate the 6SK7 (Fig 2) requires a device with a transconductance of about 2000 μmhos and a dynamic output resistance greater than 800 kΩ.[1] The g_{fs} of a JFET can be expressed as:

$$g_{fs} = \frac{2I_D}{(V_{off} - V_{gs})} \qquad \text{Eq 2}$$

where I_D is the drain current, V_{gs} is the gate-to-source voltage that causes I_D, and V_{off} is the value of V_{gs} that reduces I_D to zero.

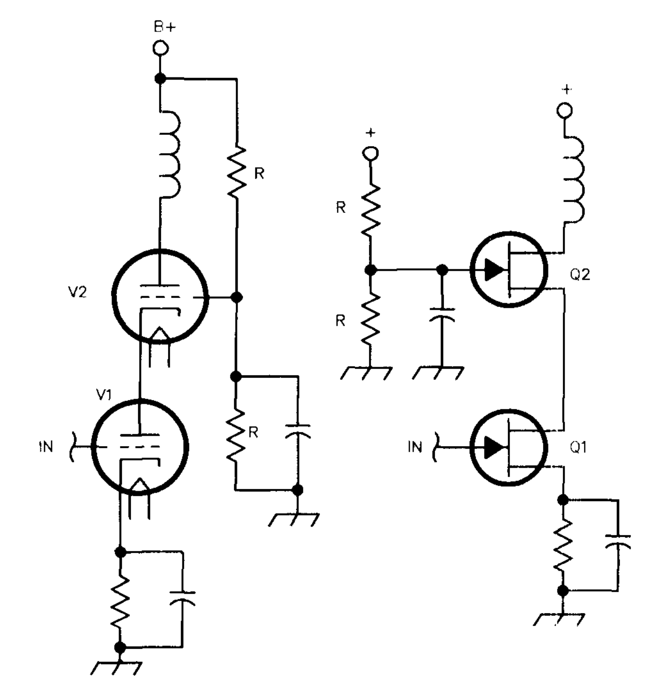

Fig 1—The cascode uses triodes as an RF amplifier.

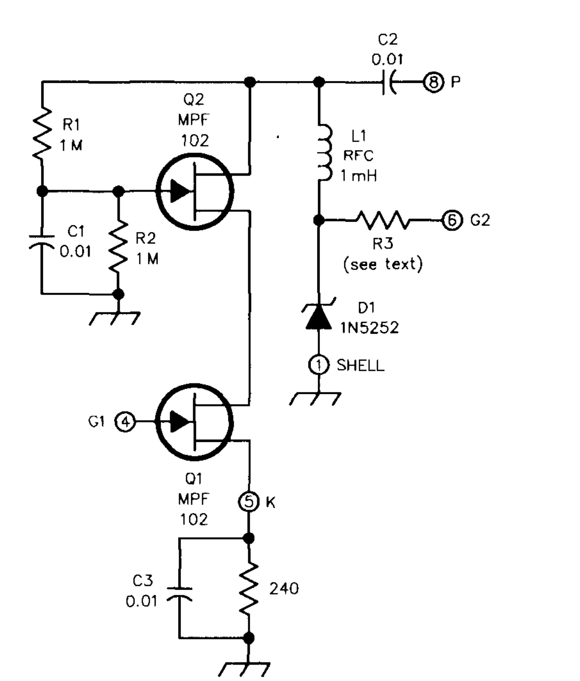

Fig 2—A JFET cascode synthesizes a 6SK7 pentode.

The fundamental electrical parameters of a JFET, I_D, V_{off} and V_{gs} are related as follows:

$$I_D = I_{DSS}\left(1 - \frac{V_{off}}{V_{gs}}\right)^2 \qquad \text{Eq 3}$$

where I_{DSS} is the drain current when V_{gs} is 0. Eq 1 can be rewritten to solve for V_{off} or V_{gs}:

$$V_{gs}/V_{off} = 1 - \sqrt{I_D/I_{DSS}} \qquad \text{Eq 4}$$

$$V_{off} = V_{gs}/\left(1 - \sqrt{I_D/I_{DSS}}\right) \qquad \text{Eq 5}$$

$$V_{gs} = V_{off}\left(1 - \sqrt{I_D/I_{DSS}}\right) \qquad \text{Eq 6}$$

The values of V_{off}, V_{gs} and I_{DSS} given in the data sheets are often only maximum and minimum values and are quite broad. For circuit design, either typical values are assumed or actual values are measured.

The actual values can be measured with simple test equipment: a power supply (something between 9 and 24 V), a multimeter and a resistor in the range of 20 kΩ. Connect the N-channel transistor's drain to the positive side of the power supply, the gate to the negative side of the power supply, and the source to the negative side through the resistor. Measure the voltage E across the resistor R and calculate or measure I_D. E is V_{gs}. Short the resistor and measure I_{DSS}. Measured values for a particular MPF102 were found to be: I_{DSS} = 4 mA, V_{gs} = 2.62 V for I_D = 0.131 mA. Plugging these values into Eq 5 and solving for V_{off} yields V_{off} = 3.2 V.

The value of g_{fs} corresponding to I_D and I_{DSS} can be found by substituting Eq 6 into Eq 3:

$$g_{fs} = \frac{2\sqrt{I_D I_{DSS}}}{V_{off}}$$

Rewriting to solve for I_D yields:

$$I_D = \frac{\left(g_{fs} V_{off}\right)^2}{4 I_{DSS}} \qquad \text{Eq 7}$$

Eq 7 shows that g_{fs}, the parameter we're trying to synthesize, is related to I_D. Eq 6 shows that I_D is related to V_{gs}. Thus if we establish the proper value of V_{gs}, we should get the value of g_{fs} we want. For the typical 6SK7 g_m of 2000 µmhos, Eq 7 shows that the I_D that produces a V_{gs} of 2000 µmhos in the MPF102 measured earlier is 2.56 mA. Using Eq 6, the V_{gs} corresponding to that I_D of 2.56 mA is calculated to be 0.64 V. For this particular JFET, a 0.64-V V_{gs} should produce the desired g_{fs} of 2000 µmhos.

A 242-Ω source resistor will produce this V_{gs}. A standard 240-Ω part is used. The typical cathode bias for a 6SK7 used in an IF amplifier is 3 V, which is provided by 270 Ω bypassed with a 0.1-µF capacitor. 270 Ω is pretty close to 242 Ω and may be used with only a moderate change in performance, within the limits of the tube. It is serendipitous that the same value of bias resistor can be used for either the FET or the tube. A different FET would probably require a different resistor.

The gain of the synthesized tube is probably the most important characteristic to be developed, but the input and output capacitances can be important, too. The input capacitance of a common-source amplifier is greater than the gate-to-source capacity by virtue of the *Miller effect*:

$$C_{in} = C_{iss} + C_{rss}|A_v|$$

where C_{iss} is the capacitance from gate

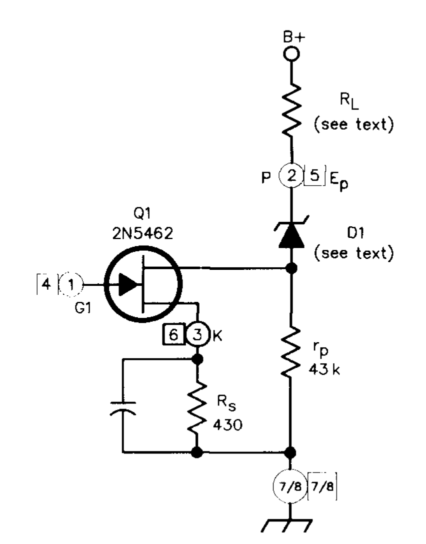

Fig 3—A JFET synthesizes one half of a 6SL7.

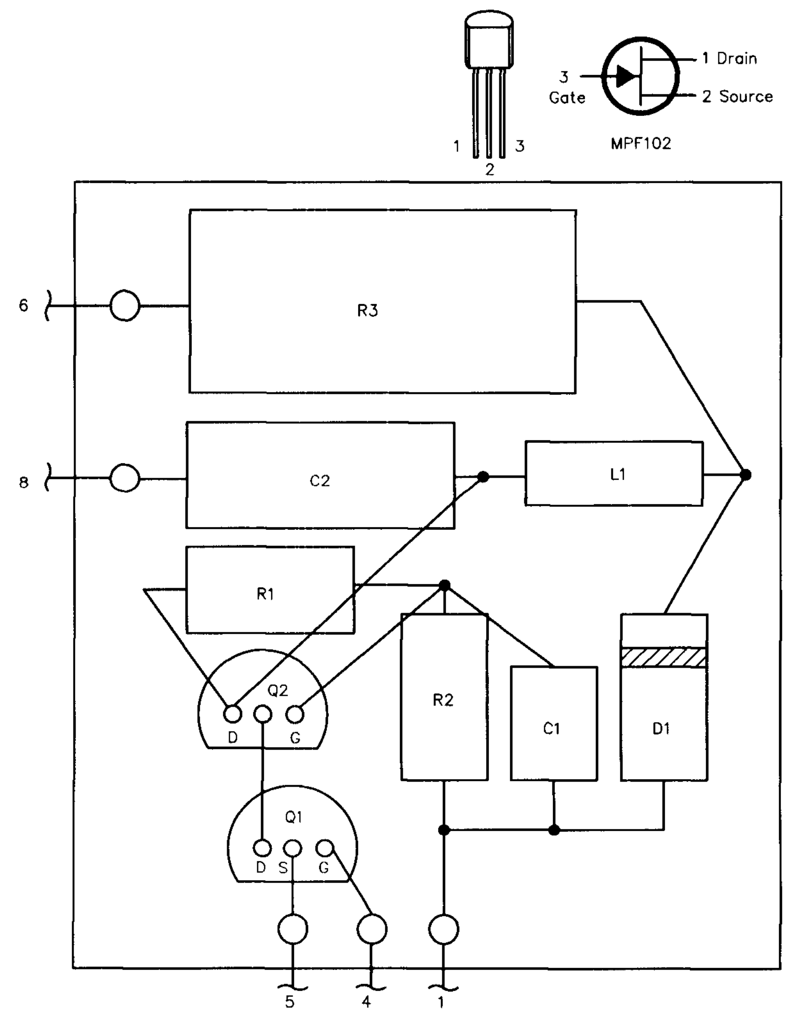

Fig 4—A perf-board layout for a synthesized 6SK7.

to source, C_{rss} is the capacitance from gate to drain, and A_v is the voltage gain of the amplifier. In a cascode amplifier, the voltage gain of the common-source part is essentially −1 (the input is inverted), so the input capacitance is simply $C_{iss} + C_{rss}$. For an MPF102, that's about 10 pF. The capacitances of a 6SK7 are: C_{in} = 6 pF, C_{out} = 7 pF, C_{gp} = 3 fF. The differences can undoubtedly be accommodated by the coupling transformer tuning. Low output-to-input capacitance C_{gp} is critical for stable RF/IF gain. Both the pentode and the FET cascode have feedback capacitance in the femtofarad range. But from a practical point of view, a stable amplifier is as dependent on the physical layout, interstage shielding and dimensions of the chassis as on the plate-to-grid capacitance.

Practical Replacement Circuits

Fig 2 shows a cascode JFET replacement for the 6SK7 pentode. The only sticky part of using JFETs to synthesize vacuum tubes is obtaining the needed low operating voltages with minimum modification of the receiver. The dc plate voltage of the 6SK7 is typically 150 V, while the maximum drain-to-source voltage of the MPF102 is specified as +25 V. The gate voltage of Q2 sets the source voltage of Q2, which is the drain voltage of Q1. The resistors R1 and R2 in the gate divide the voltage to about half the drain voltage, so a supply of 50 V would result in about 25 V across each of Q1 and Q2. Using a nominal 43-V supply allows for tolerances in R1 and R2 as well as for up to 10% variation in the supply voltage. The lower drain-supply voltage needed for the FETs can be obtained in several ways. An obvious solution is to build a suitable low-voltage supply, but that would be a major receiver modification that would be hard to justify unless all of the tubes were replaced.

Conceivably, the plate voltage could be dropped using a bypassed resistor. The specifics of dropping the voltage would depend on the drain current and the B+ supply voltage used in the receiver. The dropping resistor can be calculated, and the drain voltage is $V_{drain} = B^+ - RI_D$. Since I_D is constant, the voltage across R is constant, and B+ changes appear at the drain. Most tube-type receivers have an unregulated plate supply, and line voltage changes of ±10% cause a 150-V supply voltage to change ±15 V. The power supply voltage fluctuations can thus drastically change the FET's supply voltage. A simple dropping resistor in the high-voltage plate supply may be a risky solution.

A more conservative solution for RF/IF amplifiers is to regulate the screen supply down with a Zener diode and shunt feed the JFETs through an RF choke, as shown in Fig 2. (The screen supply is not otherwise needed.) The supply voltage for the cascode can be anything between 9 and 47 V. The Zener used in this example is arbitrarily chosen to be 24 V. The power dissipated in the dropping resistor, R3, is $(B^+ - V_Z)(I_D + I_Z)$ and may be in the 1/2-W range. It is good practice to derate resistor power dissipation by 50%, so a 1-W part would be appropriate for R3. A 1-mH choke, similar to a J. W. Miller part number 9230-92, is adequate for L1. The coupling capacitor, C2, must have a voltage rating greater than the B+ supply; use something similar to a Sprague 5HKSS10. The gate-bypass capacitor, C1, must have a dc working voltage greater than half the drain voltage; a monolithic ceramic like Sprague's 1C10X7R103K100B rated at 100 VDCW is a convenient size.

The bias will probably need to be changed from the pentode tube's value. Changing the bias means changing the value of the cathode resistor, but keeping the cathode bypass capacitor.

Almost any pentode can be synthesized with the equations given and can be substituted for hard to find tubes. The layout of the circuit board is not especially critical. A perf board mounted on an octal header will not take up any more room than the tube and can be pin-for-pin compatible with the tube. A layout of the perf board is shown in Fig 4. The terminal numbers shown are those corresponding to a 6SK7. The shell of the tube connects to pin 1, which should go to ground. The positive supply for the FETs comes through the G2 pin, pin 6. The G2 supply, V^+, is typically 100 V and may come from the plate supply through a dropping resistor or from a separate supply. The value of R3 needed to drop the screen voltage to the Zener voltage can be calculated with the following equation:

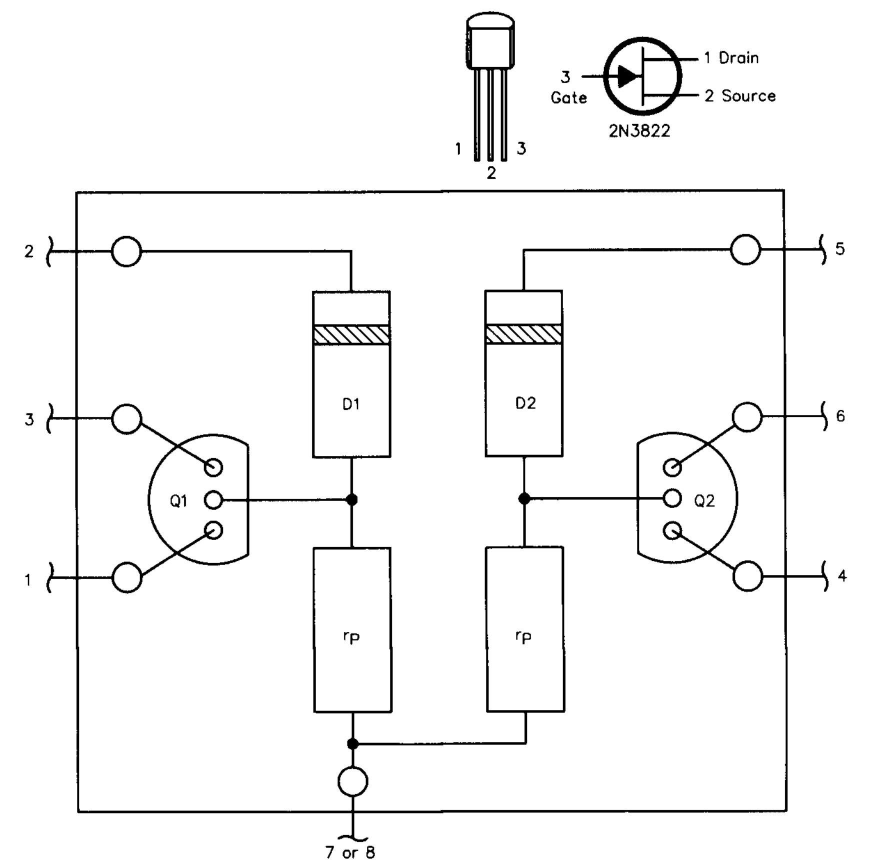

Fig 5—A board layout for a synthesized 6SL7.

$$R3 = \frac{V^+ - V_Z}{I} \quad \text{Eq 8}$$

where V^+ is the supply voltage, V_Z is the Zener voltage and I is the current drawn from V^+.

The existing screen voltage must be measured and the current limiting resistance calculated using Eq 8. The current in R3 is the sum of the drain current and the Zener current. (The Zener current need not be greater than 1 or 2 mA.)

Pin 5, the cathode pin of the 6SK7, probably goes to ground through about 270 Ω. In the example cited, the cathode resistor should be 240 Ω, but for other FETs a different value of resistance may be needed.

Synthesizing a dual-triode, such as the 6SL7, in a resistance-coupled amplifier presents a different problem: enough pins to bring in ground and the high capacitance of coupling and bypassing capacitors. The six pins needed for the two sets of triode elements and two pins for the heater occupy all eight pins of the socket. The synthesized 6SL7 doesn't require a heater voltage, so one of those pins can be used to bring in ground. The technique of applying power to the FET with shunt feed as used with the pentode is not an attractive option because high-voltage coupling capacitors for low-frequency circuits are physically large. The only other choice is direct coupling and accepting the power supply changes. In Fig 3, the supply voltage is shown being series-fed through the plate load resistor R_L, with a heater pin being used for ground.

The 6SL7 data sheet shows $\mu = 70$, $r_p = 44$ kΩ, and $G_m = 1600$ µmhos. A MPF3822 JFET is chosen to simulate the triode to take advantage of its higher (than the MPF102) drain-source voltage rating and to demonstrate handling a different JFET.

The MPF3822 has only its maximum and minimum characteristics specified, so the following values were measured: $I_{DSS} = 6$ mA, $I_D = 2$ mA for $V_{gs} = 3.5$ V. The value of V_{off} is found to be 4.3 V with Eq 5. Eq 1 shows that I_D is 2 mA when G_{fs} is 1600 µmhos. The V_{gs} needed for $I_D = 2$ mA is shown by Eq 6 to be 1.8 V. The source resistance, R2, needed to produce the 1.8 V bias is 910 Ω. The equivalent r_p is obtained by shunting the FET drain with 43 kΩ (the nearest standard 5% part). The voltage, E_p, can be shown to be:

$$E_p = \left(B^+ - R_L I_D\right) r_p \big/ \left(r_p + R_L\right) \quad \text{Eq 9}$$

B+ and R1 are peculiar to a particular receiver's circuit and must be measured. As an example, assume the supply is +150 V ±10% (135 V to 165 V), R_L is 20 kΩ and I_D is 2 mA. For these conditions, E_p varies from 65 V to 85 V. A 47-V Zener will drop E_p to 18 V to 38 V for the drain. If R_L were 47 kΩ, E_p would vary from 19 V to 34 V and no Zener would be necessary. Power dissipation in the 43-kΩ resistor is less than 1/4 W.

Conclusion

I've shown that a pair of JFETs in a cascode circuit can synthesize a pentode, and a single JFET can synthesize a triode. These rather simple circuits can solve the problem of finding tubes to keep the receiver operating, and the modifications to the receiver are nominal. The equations given above allow the scrounger of flea markets, yard sales and hamfests to resurrect a receiver that can be had for just a few dollars and some research to find out what tubes are missing. The older *Radio Amateur Handbooks* have summaries of receiving tube base diagrams and typical operating characteristics. Tube-type receivers don't have the pizzazz of the latest and greatest solid-state receivers, but they don't have the big price tag, either. Ham radio doesn't have to be out of reach of the ham who has to watch his pennies. □□

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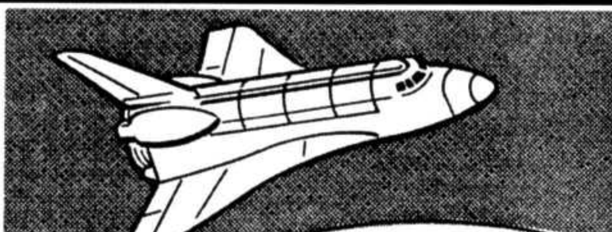

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Upcoming Technical Conferences

Eastern States VHF/UHF Conference

The 23rd Eastern States VHF/UHF Conference will be held August 22-24, 1997, at the Harley Hotel, Enfield, Connecticut.

For information on this year's conference please contact: Fred Stefanik, N1DPM, tel: 413-569-0116, ext 211; Stan Hilinski, KA1ZE, tel: 860-649-3258; or Ron Klimas, WZ1V, tel: 860-768-4758.

Microwave Update '97

Microwave Update '97 will be held October 23-26, 1997, at the Holiday Inn Conference Center in Sandusky, Ohio.

A "Surplus Tour" is scheduled for Thursday; conference papers will be presented on Friday and Saturday; noise-figure measurements and a microwave flea market are planned for Friday night; Saturday night dinner will be a Bar-B-Q; and Sunday, conference wrap up and possible tour of the W9JK "Big Ear" at Ohio State University.

Registration before October 2, 1997 is $40; after October 2, 1997 it's $45. Conference fee includes one copy of the Proceedings. Additional copies are $10. Saturday night BBQ Dinner is $15.

Hotel rates at the Holiday Conference Center are: Single, $69.95 per night; Double, $95.90. Price includes buffet breakfast and lunch on Friday and Saturday and breakfast on Thursday and Friday.

A ladies program will be offered. Planned activities include sightseeing on the north coast area (Lake Erie) and shopping. For information on area events, contact the Lake Erie Visitor's Bureau at 1-800-255-ERIE, or check their Web page at **http://www.buckeyenorth.com/**.

For more conference information contact: Tom Whitted, 4641 Port Clinton East Road, Port Clinton, OH 43452; tel: 419-732-2944.

ARRL and TAPR Digital Communications Conference

The 16th Annual ARRL and TAPR Digital Communications Conference will be held October 10-12, 1997, at the Holiday Inn BWI Airport, Baltimore, Maryland.

This year's local host is the Amateur Radio Research and Development Corporation (AMRAD).

Call for Papers: Papers for inclusion in the proceedings are due by August 20, 1997. They should be sent to Maty Weinberg at ARRL HQ.

This is a conference for all, beginners to digital experts. Topics include: APRS, satellite communications, TCP/IP, digital radio, spread spectrum and more.

Friday will include an all-day symposium covering APRS; late Friday afternoon there will be a half-day seminar entitled "RF Basics for Computer Weenies: Helping the RF-challenged get the most out of the new high-speed wireless toys"; papers will be presented on Saturday; a seminar on "Spread Spectrum System Design and Theory" will be conducted on Sunday morning.

A block of rooms have been reserved at the Holiday Inn BWI Airport at the special rate of $89 per night. The rates are good on reservations made before September 9, 1997. (Rooms cannot be guaranteed after that date.) For reservation call the Holiday Inn BWI Airport at: tel: 410-859-8400 or fax: 410-684-6778. Ask for the Digital Conference rate.

Preregistration before September 10, 1997, is $42, after September 10, or at the door it's $47. Registration includes one copy of the Conference Proceedings, sessions, meetings and lunch on Saturday. Saturday dinner: $20. Seminars/Symposiums—Friday, APRS 1-8pm: $25; Friday, RF Basics for Computer Weenies 3-7pm: $20; Sunday, Spread Spectrum System Design and Theory 8:30 am-1:30 pm: $20.

To register for the conference, or for more information, contact TAPR at: 8987-309 E Tanque Verde Road #337, Tucson, AZ 85749, tel: 817-383-0000, fax: 817-566-2544 or Internet: **http://www.tapr.org/dcc/**.