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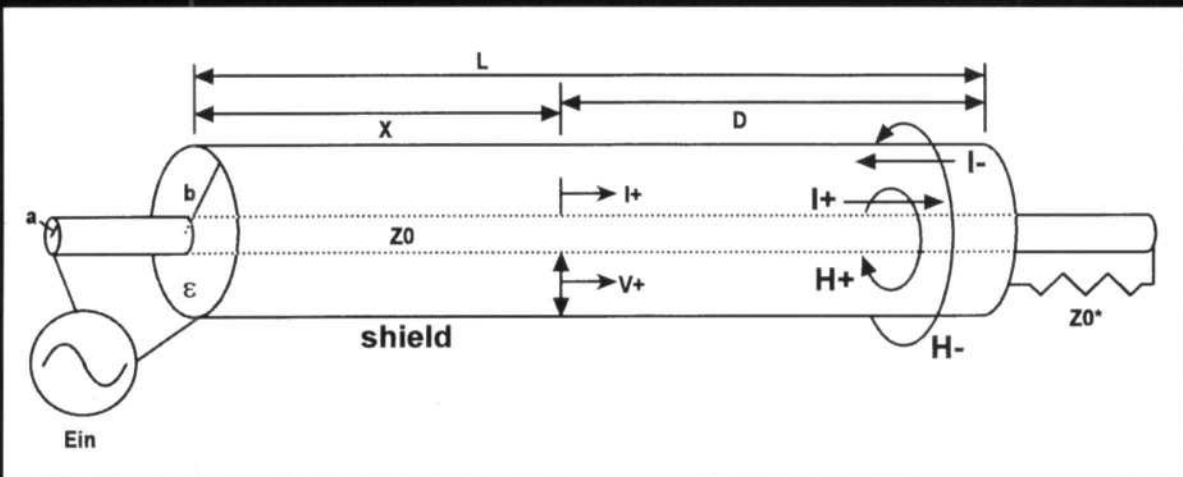
ARRL Experimenter's Exchange

August 1996




## Coax Models by Computer

**QEX**: The ARRL
Experimenter's Exchange
American Radio Relay League
225 Main Street
Newington, CT USA 06111

QEX

*QEX* (ISSN: 0886-8093 USPS 011-424) is published monthly by the American Radio Relay League, Newington, CT USA.

Second-class postage paid at Hartford, Connecticut and additional mailing offices.

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Subscription rate for 12 issues:

In the US: ARRL Member $15, nonmember $27;

US, Canada and Mexico by First Class Mail: ARRL Member $28, nonmember $40;

Elsewhere by Surface Mail (4-8 week delivery): ARRL Member $20, nonmember $32;

Elsewhere by Airmail: ARRL Member $48, nonmember $60.

*QEX* subscription orders, changes of address, and reports of missing or damaged copies may be marked: *QEX* Circulation. Postmaster: Form 3579 requested. Send change of address to: American Radio Relay League, 225 Main St, Newington, CT 06111-1494.

Members are asked to include their membership control number or a label from their *QST* wrapper when applying.

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## About the Cover

**WØIYH explains the math needed to computer-model coaxial cable.**

**ISSUE NO. 175**

# Features



# Columns



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**Purpose of *QEX*:**

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*.



# *Empirically Speaking*

## Even Paranoids Have Enemies

Recently, we at ARRL established a form-based interface on our Web server that allows you to join ARRL or renew your membership via the World Wide Web. (It's at **http://www.arrl.org/forms/member_general.html**, in case you've been putting off that renewal.) One of the issues we had to resolve was whether we could usefully do that without having a Web server capable of secure transfers, seeing as how we wanted to accept credit-card numbers via the form. In the end, we decided to go ahead with the form, offering the user the option of supplying a credit-card number or supplying a telephone number that we could use to call back to get the credit-card information.

We had no idea whether most people would be willing to have their credit-card information transmitted unencrypted across the Internet. Interestingly, it turns out that most people are willing to do so.

This really makes perfect sense. While it is possible under the right circumstances for a third party to intercept information of that sort, it isn't common. To put the risk in perspective, we note that the local news here in Connecticut recently reported that a number of credit-card transaction slips were found in a dumpster behind a local business. Anyone care to speculate on how often that happens compared to interception of unencrypted credit-card numbers on the Internet?

Even so, there are plenty of people who report that they will *never* send their credit-card information via unencrypted Internet transfers. It's hard to avoid thinking that the concern about this subject is largely a result of media frenzy. A lot of articles have appeared in the general media about Internet security, and some have been worthy of appearance in the weekly supermarket tabloids. Still, from the perspective of an organization wishing to provide credit-card ordering of services, it's not what *we* think that matters, it's what the *users* think.

So, we'll soon be running a secure Web server at ARRL HQ. There are enough people who want to see it done that way to make it worth the doing, so we will. It's a funny place, the Web.

## *CyberHam* Caters to the Computerized Ham

We recently saw the premiere issue of *CyberHam* magazine, published by Harlan Technologies, 5931 Alma Drive, Rockford, IL 61108 (gharlan@cris.com). *CyberHam* promises to combine computers and Amateur Radio. The initial issue includes articles covering aspects of Amateur Radio and the Internet and Compuserve, Linux, packet radio, satellites and other topics. Publisher Gene Harlan promises great things to come; we'll be watching with interest!

## This Month in *QEX*

It's one thing to toss a chunk of coax between your transmitter and antenna and get the watts where you want them to go. It's quite another to actually understand how that happens and to apply that understanding to more critical applications. The latter approach can be helped by "Computer Modeling of Coax Cable Circuits," as described by William E. Sabin, WØIYH.

Packet voice: it's an idea that has been discussed and dreamed about for quite a while. If you'd like to experiment with a simple, inexpensive project, try "A Packet Voice Communication System," by Dirench Dogruoz, Carlos Lopez, Harvind Samra, KBØPRP, and Brett Weldy.

You've seen those HF mobile antennas with the big capacitance hats atop them. Maybe you've even built one or two. But did you ever try to design an antenna using one? If so, you probably performed a lot of cut-and-try. Andrew S. Griffith, W4ULD, can help you reduce the cut-and-try in "Capacitance Hats for HF Mobile Antennas."

Peter Traneus Anderson, KC1HR, keeps finding better analog-to-digital converters to use with his direct-digital-conversion receiver. Catch the latest in "A Better and Simpler A/D for the DDC-Based Receiver."

Lastly, we provide a list of 1996 conference proceedinmgs available through ARRL.—*KE3Z, email:jbloom@arrl.org.*

# Computer Modeling of Coax Cable Circuits

*Why fool around with approximations and guesses? You can know exactly what's happening on your coaxial cable.*

By William E. Sabin WØIYH

This article will try to bring under one roof the information that is needed to understand the behavior of, and calculate the performance of, ordinary types of coax cable that we commonly use in amateur-radio work. It's been my experience that this information resides in bits and pieces in numerous articles, handbooks and textbooks, but is somewhat tedious and confusing to pull together into a clear and simple overview that's easy to use to understand and design RF systems. The approach is to put you in direct contact with mathematical formulas that can be easily evaluated on a personal computer, for example in a *Mathcad* worksheet.[1] The results of the calculations can then often be used by the *ARRL Radio Designer* program and other software to perform simulations and optimizations of various kinds. I'll try to keep the mathematics fairly, but not perfectly, painless. For many, much of this material will be a review.

[1]Notes appear on page 10.

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### The Basic Idea of Coax

A length of coax conveys RF power. A step voltage applied to one end creates a power wave that reaches the load at some later time. Some of this power may return back to the sending end at a still later time. The two conductors (the center lead and the braid) are said to act as *guides* for this power wave,[2] and that the electromagnetic energy is actually conveyed, in a mathematical sense, by the electric and magnetic fields that exist within the dielectric material. After the initial, or *transient* conditions, have all been resolved, this traveling wave viewpoint is no longer mandatory and we find that in the *steady state* a length $l$ of coax is a device that *transforms* impedances, voltages and currents in a way that is similar to, but not exactly like, a conventional transformer. We will focus on these steady-state conditions.

The key to understanding coax behavior lies in understanding the electric and magnetic fields. For example, the current flow in the center lead creates a magnetic field that encloses the current. The return current in the braid also creates a magnetic field that lies *entirely outside* the region of its current flow. These two flux fields cancel each other on the *outside* of the coax. The electric field exists only *between* the conductors. These are the principal shielding mechanisms that keep the signal inside the coax. Various imperfections detract from this goal, and the coax braid is not a perfect shield. An excellent reference on coax leakage is listed in Note 3. The braid also protects, to a large extent, the inside of the coax from external electric and magnetic fields. The exact electromagnetic theory principles by which these and many other effects occur are complicated and are beyond the scope of this article.

At radio frequencies, a segment of coax can function as a *conventional* transformer and is often used this way. The center lead can be a primary and the braid a secondary (and vice versa). The turns ratio in this mode is 1:1.

### Characteristic Impedance

Fig 1 shows a length of coax connected to an impedance $Z_0^*$ (the reason for this will become apparent later) that equals the complex conjugate of $Z_0$, the *dynamic characteristic impedance* of the cable. This means that as a power wave travels toward the load, the ratio of the voltage $V^+$ to the current $I^+$ is $Z_0$. At *any* frequency this ratio is:

$$Z_0 = \frac{V^+}{I^+} = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \quad \Omega \qquad \text{Eq 1}$$

where $R$ is the series resistance ($\Omega$) per meter and $L$ is the series inductance (H) per meter. $G$ is the shunt conductance (S) per meter and $C$ is the shunt capacitance (F) per meter of the polyethylene dielectric material. $R$ and $G$ are responsible for the power losses (attenuation) in the cable and also for a reactive component to $Z_0$.

This formula is meaningful to us only at radio frequencies where:

$$\omega L >> R \text{ and } \omega C >> G \qquad \text{Eq 2}$$

and as frequency $\omega(=2\pi f)$ increases, $Z_0$ would seem to be getting closer and closer to the simple approximate pure resistance value, which is nearly always used in most ordinary low-loss applications:

$$Z_0 \approx \sqrt{\frac{L}{C}} = R_0 \qquad \text{Eq 3}$$

but it turns out that $R$ increases as the square root of frequency and $G$ increases directly with frequency (more about this later), so that this ideal value is somewhat elusive.

In some situations, when the attenuation is not negligible but also not large, we would like to have a more accurate value of $Z_0$:

$$Z_0 \approx R_0 \left[1 - j\left(\frac{R}{2\omega L} - \frac{G}{2\omega C}\right)\right] \qquad \text{Eq 4}$$

and we see more clearly how the cable losses $R$ and $G$ cause $Z_0$ to be slightly reactive, which is a rather peculiar result. At mid frequencies the term $G/2\omega C$ is usually much less than $R/2\omega L$ (dielectric losses are small) and Eq 4 further simplifies to:

$$Z_0 \approx R_0 \left(1 - j\frac{R}{2\omega L}\right) \qquad \text{Eq 5}$$

but at the highest UHF frequencies, because $G$ increases more rapidly than $R$, greater accuracy results if we use Eq 4. At some microwave frequency, coax (especially thick coax) starts to behave somewhat like a complicated waveguide, but we will not be concerned with that.

In this article meters and centimeters will be used, to be consistent with the prevailing literature. To convert feet and inches to meters and centimeters:

| | |
|---|---|
| m = f x 0.30480 | $f^{-1} = m^{-1}$ x 0.30480 |
| f = m x 3.2808 | $m^{-1} = f^{-1}$ x 3.2808 |
| cm = in x 2.540 | $in^{-1} = cm^{-1}$ x 2.540 |
| in = cm x 0.3937 | $cm^{-1} = in^{-1}$ x 0.3937 |

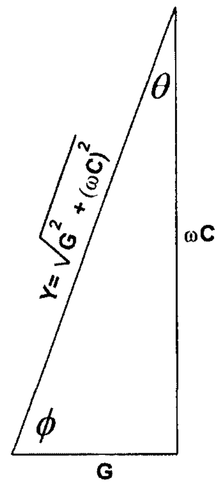


Fig 2—A graphical definition of loss tangent and power factor.

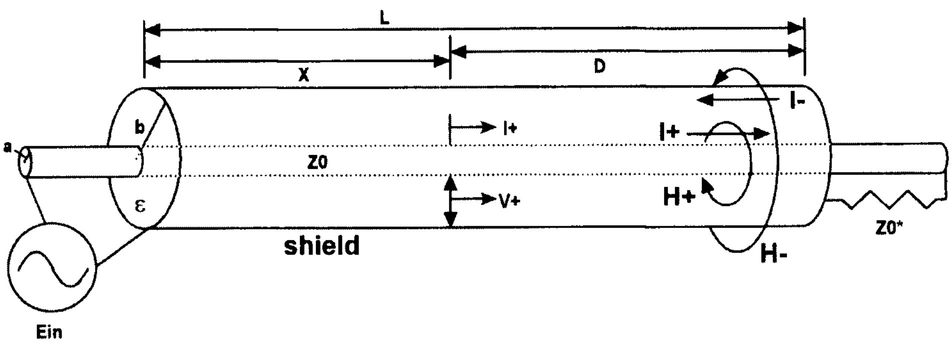


Fig 1—A coaxial cable of characteristic impedance $Z_0$ and length $L$, terminated in $Z_0^*$. The currents and H-fields are shown.

Later, we will revisit and improve these formulas as we develop additional information (see Eq 20). First, we will give some details regarding $Z_0$, $R$, $L$, $G$ and $C$ that are interesting and useful.

### Coax Cable Parameters

The inductance $L$ per meter is determined by finding the magnetic flux $\phi$ per meter of length, between the center lead and the braid, that is generated by a current $I$ in the center lead, according to Ampere's law. (Note: the return current through the braid does not generate flux in the dielectric region between the center wire and the braid.) The ratio of flux to current is, by definition, the inductance per meter ($N$=1, one single turn, is assumed):

$$L = 0.2\ln\left(\frac{b}{a}\right) = 0.00333 R_0 \sqrt{\varepsilon_r} = 0.00333 \frac{R_0}{VF} \ \mu\text{H/m} \qquad \text{Eq 6}$$

where ln is the natural logarithm. In this equation, $b$ and $a$ are shown in Fig 1, $\varepsilon_r$ is the relative dielectric constant and $VF$, the velocity factor, is the ratio of the wave's phase velocity to the speed of light (in air or vacuum). This number, for many types of 50-Ω coax, is $VF \approx 0.64$ to 0.67, with 0.66 as typical. It follows from Eq 6 that:

$$\sqrt{\varepsilon_r} = \frac{1}{VF} \qquad \text{Eq 7}$$

which makes $\varepsilon_r \approx (1/0.66)^2$ or about 2.3 for the commonly used polyethylene dielectric material. For a 50-Ω coax, Eq 6 finds the $b/a$ ratio to be:

$$\ln\left(\frac{b}{a}\right) = \frac{(0.00333)(50)}{(0.2)(0.66)} \quad \text{or} \quad \frac{b}{a} = e^{1.25} = 3.49 \qquad \text{Eq 8}$$

From Eq 6, a typical value for $L$ is $0.00333 \times 50 \div 0.66 = 0.252$ μH per meter.

The capacitance $C$ per meter is found by relating a charge in coulombs per meter on the center wire to the voltage difference between center lead and braid, according to Gauss's law. The result for coax is:

$$C = \frac{\varepsilon_r}{18\ln\left(\frac{b}{a}\right)} \bullet 10^3 = \frac{\sqrt{\varepsilon_r}}{3R_0} \bullet 10^4 \quad \text{pF/m} \qquad \text{Eq 9}$$

A typical value of $C$ for 50-Ω coax is approximately 101 pF per meter. Using these typical $L$ and $C$ values, the square root of 0.2525 μH / 101.1 pF gives a $Z_0$ of 50.0 Ω. Combining Eqs 6 and 9, we can now rewrite Eq 3 as follows:

$$Z_0 \approx \sqrt{\frac{L}{C}} = \frac{60}{\sqrt{\varepsilon_r}}\ln\left(\frac{b}{a}\right) = \frac{138}{\sqrt{\varepsilon_r}}\log_{10}\left(\frac{b}{a}\right) = R_0 \quad \Omega \qquad \text{Eq 10}$$

The dielectric material of the cable has a capacitive susceptance $j\omega C$ per meter and is in parallel with a conductance $G$ per meter. In Fig 2 the very small angle $\theta$ between $Y$ and $\omega C$ is the *loss angle* and the tangent of that angle, $G/\omega C$, is the *loss tangent* $T_L$. A small angle implies a low value of $G$ and for this small angle the loss tangent is practically the same as the *power factor* $F_P$, which is $\cos(\varphi) = G/Y$, which is also used. These are specified for various dielectric materials (for polyethylene $\approx 5 \times 10^{-4}$) and so:

$$G = T_L \bullet \omega C \approx F_P \bullet \omega C \qquad \text{Eq 11}$$

As an example, for 1 meter of 50-Ω cable at 100 MHz $G \approx 3 \times 10^{-5}$ S. The dielectric losses are due mostly to hysteresis effects in the dielectric molecules as the electric field oscillates.

The resistance per meter, $R$, involves the center lead and the braid, with the resistance of the center lead being typically four times greater.[4] The basic resistivity of copper is $1.75 \times 10^{-8}$ Ω-meter, which means that a cube of copper, one meter to a side, would have that value of resistance between two opposite faces of the cube. The formula for the resistance of any cube is:

$$R_{\text{cube}} = \frac{\rho}{l} \qquad \text{Eq 12}$$

where $\rho$ is resistivity and $l$ is the length of one side. A cube of 1 cm per side has 100 times as much resistance as a 1-meter cube.

As frequency increases the resistance increases due to skin effect. That is, the current in the center conductor flows mostly along the outer surface. If the wire is tin-plated, a significant part of the current flows in the tin, whose resistivity is 6.7 times greater than copper. So low-loss coax avoids tin plating, despite its presumably easier soldering property (RG-11 is an example). If the center lead is stranded the resistance is about 1.3 times higher because of the spiraling of the strands and the contact resistance between adjacent strands, so a solid wire is preferred, but stranded is often used anyway because it is more flexible. Increasing the size of the center lead reduces its loss significantly but also reduces $Z_0$, according to Eq 10. To maintain the same $Z_0$ the cable diameter must be increased. If, instead, a foam dielectric with a lower $\varepsilon_r$=1.56 is combined with a larger conductor, $Z_0$ and the cable diameter can be maintained at the same size, but with lower copper loss. RG-8 foam ($VF$=0.80) is an example.

The braided shield also contributes to resistance. The current flows on the inner surface, also because of skin effect. The manner in which the wire is braided is important, and the outer jacket presses tightly against the braid to reduce resistance and improve stability under flexure. Manufacturers use *braid factors* to characterize their products.

Skin effect is a consequence of Ampere's law and Faraday's law. Fig 3 illustrates the action. The rapidly changing magnetic flux within the loop a-b-c-d-a, due to current *inside* the wire (Ampere's law), induces an electric field (Faraday's law) within this loop. This electric field is in such a direction that the current in path a-to-b is retarded and the current in path d-to-c is enhanced. The induced electric field strength increases directly with frequency, forcing the current ever closer to the surface at path d-to-c. The result is that the interior of the wire is very nearly an insulator for the current in Fig 3. Fig 3 also shows E-field components from a-to-d and from c-to-b, but these are canceled out by opposing E-fields in adjacent segments to the left and right. The same action occurs on the braid,

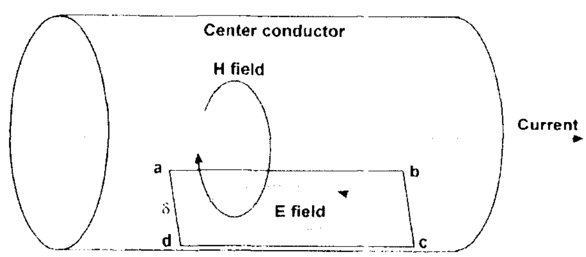


**Fig 3—The actions of the E- and H-fields are responsible for skin effect.**

forcing the current toward the inside surface of the braid.

At a distance $\delta$ (the skin depth) from the surface the current density has reduced to the factor $e^{-1} = 0.37$ (37%) and at $5\delta$ it is $e^{-5} = 0.0067$ (0.67%), where $e = 2.7183$. The value of $\delta$ for copper is:

$$\delta = \sqrt{\frac{\rho}{\pi f \mu}} = \frac{6.64}{\sqrt{f}} \quad \text{cm} \qquad \text{Eq 13}$$

where $\mu$ is the magnetic permeability of copper ($4\pi \times 10^{-7}$ H/m), the same as free space. At 1.8 MHz, $\delta$=0.049 mm or 0.0019 in.

Because of the magnetically induced electric field *inside* the wire, there is a phase lag in the *internal* current in the amount of 1.0 radian at a depth of $\delta$. This is called the *internal inductance* effect. This inductance is *not* the same as the $L$ in Eq 1, but at low frequencies it makes a small contribution to it and it almost disappears at high frequency. For copper coax, the total resistance of the center wire plus braid is approximately (the vendor's correction factors are needed):

$$R = \frac{1}{2}\sqrt{\frac{\rho f \mu}{\pi}}\left(\frac{1}{a} + \frac{1}{b}\right) = \frac{4.17 \bullet 10^{-6}\sqrt{f}}{b}\left(\frac{b}{a} + 1\right) \quad \Omega/\text{m} \qquad \text{Eq 14}$$

where $a$ and $b$ are in *centimeters* (see Fig 1) and the resistance increases as the square root of $\rho$ and $f$. The *internal* inductance *decreases* at the same rate. The ratio $b/a$ determines $Z_0$ (see Eq 10) so for a given $Z_0$ a greater coax diameter $b$ reduces $R$.

### Wave Propagation (No Reflections)

Let's look at the behavior of the wave as it travels from the generator to the load. For starters, assume that the load impedance $Z_L = Z_0^*$ and both are almost resistive. In other words (see Eqs 4 and 5), the attenuation of the cable is very small. We are looking for the *matched* performance and there are negligible reflections from the load, as mentioned earlier.

If an ac voltage with rms value $E_s$ is applied (Fig 1) the voltage $E$ at point $x$ is:

$$E(x) = \underbrace{E_s e^{-\gamma x}}_{A} = \underbrace{E_s e^{-\sqrt{(R+j\omega L)(G+j\omega C)}\cdot x}}_{B} = E_s \underbrace{e^{-\alpha x}}_{C} \underbrace{e^{-j\beta x}}_{D};$$

$$I(x) = \frac{E(x)}{Z_0} \qquad \text{Eq 15}$$

This is the steady-state solution to a differential wave equation for a sine-wave signal that is traveling in *one* direction—toward the load. Term A contains $\gamma$, the *propagation constant* per meter. Term B shows that $\gamma^2$ is the product of the series impedance, in $\Omega$/m, and the shunt admittance, in S/m. Term C contains a, the *attenuation constant*, in nepers per meter, and term D contains $\beta$, the *phase constant,* in radians per meter. Comparing A with C and D, we see that:

$$\gamma = \alpha + j\beta \qquad \text{Eq 16}$$

In term C, when $ax = 1.0$, the voltage $E$ has attenuated by 1.0 neper to the factor $e^{-1.0} = 0.37$. In other words, the exponent has the unit nepers. For example, if $\alpha$=.0001 nepers per meter, then in 100 meters the attenuation is 0.01 neper. One neper is equivalent to $20 \log_{10} e = 8.686$ dB, so in this example the voltage attenuates 0.08686 dB. That is, in length $x$ meters the reduction in voltage is:

$$\alpha\,(\text{nepers/m}) \bullet x(\text{m}) = \alpha x\,(\text{nepers}) = 8.686 \bullet \log_{10}(\alpha x)\ (\text{dB}) \qquad \text{Eq 17}$$

Since $Z_0$ is constant, this is also the reduction in power. In coax tables, the matched-line attenuation at some frequency is very often stated in dB per 100 feet. Multiplying this value by 0.032808 gives the dB per meter. Dividing this by 8.686 gives nepers per meter. So in $x$ meters the attenuation is $\alpha x$ nepers. This is called the *matched loss*, ML, if the load is the same as $Z_0$ (no reflections and therefore no standing waves).

The D term is a unit vector that rotates clockwise as $x$ increases. The phase at $x$ *lags* the phase at $x$=0. When $\beta x$=1.0 the phase has shifted −1.0 radians (−57.3 degrees) and one complete revolution is $-2\pi$ radians (−360 degrees). The total phase shift in radians is the radians per meter times the length in meters. The conversion from radians to degrees, or degrees to radians, is:

$$\text{degrees} = \text{radians} \bullet \frac{360}{2\pi} \qquad \text{Eq 18}$$

A further comparison between part B and parts C and D leads to the following close approximation equations for $\alpha$ and $\beta$, for a low loss *coax* cable:

$$\alpha \approx \frac{R}{2Z_0} + \frac{GZ_0}{2} \quad \text{nepers per meter}$$

$$\beta \approx \omega\sqrt{LC} \approx \frac{f}{VF} \bullet 2.092 \bullet 10^{-8} \quad \text{radians per meter}$$

$$= \frac{2\pi}{\text{wavelength (meters)}} = \frac{2\pi f\,(\text{MHz})}{\text{VF} \bullet 299.79} \quad \text{radians per meter}$$

Eq 19

If $G$ is negligible, we can find $R$, in $\Omega$/m, if the attenuation in nepers per meter is known. The formula for $\beta$, in radians per meter, is obtained by combining Eq 19 with Eqs 7, 9 and 21, and we see that $\beta$ increases as frequency $f$ (Hz) increases, and also as VF gets smaller.

Using the easily determined values of $\alpha$ and $\beta$, we can restate the low loss value for $Z_0$ that was given in Eq 5.

$$Z_0 \approx R_0\left(1 - j\frac{R}{\omega L}\right) = R_0\left(1 - j\frac{2\alpha R_0}{2\beta R_0}\right) = R_0\left(1 - j\frac{\alpha}{\beta}\right) \qquad \text{Eq 20}$$

Another frequently needed number is the wavelength, $\lambda$, in meters or feet, of a length of coax, which is the free space wavelength times the velocity factor.

$$\lambda\,(\text{meters}) = \underbrace{\frac{2.998 \bullet 10^8\ (\text{m/s})}{f(\text{Hz})}}_{\text{free space wavelength } \lambda_0} \bullet \text{VF}$$

Eq 21

$$\lambda\,(\text{feet}) = \frac{983.6}{f(\text{MHz})} \bullet \text{VF}$$

### *ARRL Radio Designer* Program

Various handbooks such as the *ARRL Antenna Book*, the *ARRL Handbook* and the *ITT Reference Data for Radio Engineers* contain graphs of $\alpha$, in dB per 100 feet, on a logarithmic vertical axis, versus frequency on a logarithmic horizontal axis, for a large number of coax types. On this kind of log-log graph the plots are approximately straight-line. Also, the transmission line models in simulation programs such as ARRL's *Radio Designer* (related to Compact's *Harmonica*) and others have options to include the frequency variation of a per unit length over a frequency range. In the two examples cited the form is:

$$\alpha\,(\text{dB/unit length}) = P\left(C1 \bullet \sqrt{f} + C2 \bullet f\right) \qquad \text{Eq 22}$$

where $P$ is the number of unit lengths (meters, hundreds of feet, etc), $C1$ relates to resistance loss and $C2$ relates to dielectric loss. The values of $C1$ and $C2$ for a particular coax can be closely approximated by *curve-fitting* a line to the graph. The *Mathcad* graph plotting feature is very useful for this task. We advise the consistent use of meters of length.

The ARRL *Radio Designer* software is a valuable resource for analyzing transmission line circuits. It can be used to get frequency sweeps of complete systems, for example a transmatch, transmission line and a set of antenna impedance values. It can perform optimization and tuning operations of many kinds.[5] This program collaborates with *Mathcad*-type programs to give us some very powerful capabilities.

**Wave Propagation (With Reflections)**

If the load impedance $Z_L$ is not equal to the complex conjugate of the complex $Z_0$, a portion of the *forward* power wave $(V^+)^2/Z_0$ is said to be *reflected* from the load back toward the generator. An equivalent statement is that if the load impedance $Z_L$ is not the complex conjugate of the complex $Z_0$, an impedance mismatch occurs and there is a *mismatch loss*, exactly the same as in lumped circuit theory, and the power that is accepted by the load is less than the power that is available. For the power wave, the difference between the forward power wave arriving *at* the load and the reflected power wave that is returned *from* the load is the power that is dissipated *in* the load, and it, plus the power that is lost in the line itself, is equal to the power that is *delivered* to the line by the generator. That is:

$$P_{\text{generator}} = \overbrace{\underbrace{P_{\text{forward}}}_{\text{at the load}} - \underbrace{P_{\text{reflected}}}_{\text{from the load}}}^{\text{in the load}} + P_{\text{line loss}} \qquad \text{Eq 23}$$

The phases of the forward and reflected voltage and current waves interact, either to enhance or diminish each other in a repetitive manner along the line. In order to limit the size of this article, we will deal mainly with the mathematical descriptions of this behavior, and avoid the lengthy word descriptions that are often used. The justification for this is that, these days, sophisticated and inexpensive math software, such as *Mathcad*, removes all of the pain from solving and *plotting* the difficult complex algebra equations so that the *visual* results are immediately available. Also, we will deal only with the rms values of steady-state sine-wave signals. With this approach, the discussion will be very clear and simple.

We are concerned with low-loss cable and mostly with voltage standing wave ratios (VSWR) that are not greater than, say 30 or so (VSWR will be discussed). The reasons for this will come later and this situation describes nearly all of the practical applications that we radio amateurs encounter in well-designed systems. In certain applications, though, such as short-circuit and open-circuit stubs or highly reactive loads, extremely high VSWRs are encountered and in fact the VSWR concept itself becomes impractical, especially when $Z_0$ becomes complex (see later discussion).

**Standing Waves**

Fig 4 shows a cable with length $l$ and certain values of $Z_0$, $\alpha$ and $\beta$, connected between a generator and a complex load $Z_L$ which is different from $Z_0$. The value $x$ is the distance from the generator and $d$ is distance from the load. Waves of voltage and current are traveling in both directions. Eqs 24 through 27 are the steady-state solutions to the wave equation for these bidirectional waves. We want to find the voltage, current, power level, impedance (or admittance) and VSWR at every point on the cable, over some frequency range and (often simultaneously) over some range of load values. We also want the power into the cable, the power into the load and the total cable loss. The procedure is to let the math program do the grunt work, solving and plotting the standard equations that we present. Our goal is to show a general approach that can be applied to many specific situations.

To find the complex impedance at any distance $d$ from the load, looking toward the load:

$$Z_d = Z_0 \frac{Z_L + Z_0 \tanh \gamma d}{Z_0 + Z_L \tanh \gamma d} \qquad \text{Eq 24}$$

where $\gamma$ is found from previous equations, $Z_0$ and $\tanh \gamma d$ are complex, and if $d=l$, the length of the coax, then $Z_d = Z_{\text{in}}$, the input impedance. If $Z_L$ is a short- or open-circuit, then from Eq 24:

$$Z_{\text{sc}} = Z_0 \tanh \gamma d \ \text{(short ckt) or } Z_{\text{oc}} = \frac{Z_0}{\tanh \gamma d} \ \text{(open ckt)} \qquad \text{Eq 25}$$

If some reference value of input voltage $E_{\text{in}}$ (eg, 1.0) is assumed, the voltage at some distance $x$ (possibly $l$) from the generator is:

$$E_x = E_{\text{in}}\left(\cosh \gamma x - \frac{Z_0}{Z_{\text{in}}} \sinh \gamma x\right) \ \text{V} \qquad \text{Eq 26}$$

and the current is:

$$I_x = \frac{E_{\text{in}}}{Z_{\text{in}}}\left(\cosh \gamma x - \frac{Z_{\text{in}}}{Z_0} \sinh \gamma x\right) \ \text{A} \qquad \text{Eq 27}$$

Computer-generated plots of the magnitudes $|E_x|$ and $|I_x|$ of Eqs 26 and 27 display the *exact* standing-wave patterns for voltage and current. The power in the load is:

$$P_L = E_L \bullet E_L^* \bullet \operatorname{Re}\left(\frac{1}{Z_L}\right) = |E_L|^2 G_L \ \text{W} \qquad \text{Eq 28}$$

where Re means the real part, $E_L^*$ is the complex conjugate of $E_L^*$, $Z_L$ is complex, $1/Z_L$ is the admittance $G+jB$ and the input power is:

$$P_{\text{in}} = E_{\text{in}} \bullet E_{\text{in}}^* \bullet \operatorname{Re}\left(\frac{1}{Z_{\text{in}}}\right) = |E_{\text{in}}|^2 G_{\text{in}} \ \text{W} \qquad \text{Eq 29}$$

where $E_{\text{in}}$ and $Z_{\text{in}}$ are also complex. The cable loss in dB is:

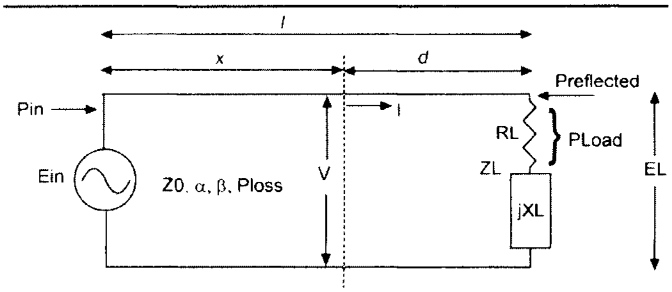


**Fig 4—Diagram of a cable and load showing the various parameters of the system.**

$$P_{loss} = 10 \log\left(\frac{P_{in}}{P_L}\right) \quad \text{dB} \qquad \text{Eq 30}$$

The voltage reflection coefficient, return loss and standing-wave ratio at the coax input are found as follows, with certain limitations that are discussed next:

$$\Gamma_{in} = \frac{Z_{in} - Z_0}{Z_{in} + Z_0} \quad \text{complex voltage reflection coefficient}$$

$$VSWR_{in} = \frac{1 + |\Gamma_{in}|}{1 - |\Gamma_{in}|} \quad \text{voltage standing wave ratio}$$

$$RL = -20 \bullet \log_{10}|\Gamma_{in}| \quad \text{return loss, dB}$$

Eq 31

Eqs 24 through 31 tell us the important things about the system, namely the input admittance of the coax (the transmatch has to deal with this), the power loss and voltages in the coax and, with some further analysis (beyond the scope of this article, but see Note 6), the voltage, current and power stresses in the transmatch (including any iron or ferrite transformers or inductors). The *input* VSWR is involved in interfacing the transmitter with the coax and must be reduced to a low number by a transmatch.[7] VSWR meters are usually calibrated to read values up to 3:1 or perhaps 10:1.

These numbers are also found for the open-circuit and short-circuit stubs and for pure reactance loads. For these situations the concepts of reflection coefficient and standing wave ratio are totally irrelevant but the real part of the input admittance enables us to find the power dissipated in the stub or line and the value of stub $Q$.

To explore this a little further, the concepts of reflection coefficient and VSWR, which are based on traveling waves of voltage or current, run into difficulty when $Z_0$ is not a pure resistance, in other words when the cable is *lossy* (see Eq 4).[8] For example, suppose that $Z_0 = 50 - j1.0$ and $Z_L = 50 + j1.0$. The line and load are then conjugately matched for maximum power transfer. But from Eq 31, the reflection coefficient is $j0.02$, the VSWR is 1.041 and RL=34 dB. For small values of reactance this error is trivial but for open-circuit or short-circuit stubs or highly reactive loads the errors get large and RL is essentially 0.0 dB. On the other hand, if we work with power values, as in Eqs 28, 29 and 30, the power numbers are correct and the usage of $\Gamma$ and VSWR is avoided and unnecessary. Also, we find that the $Z_0$ used in Eqs 24-27 can be complex, with no resultant errors (see Eq 15, part B, where the values of $L$, $C$, $R$ and $G$ imply a complex $Z_0$ as defined in Eq 1).[9] In particular, Eq 26 for the load voltage is correct for a complex $Z_0$ and therefore provides the correct value of $E_L$ in Eq 28. As a real-world matter, the error in VSWR is unimportant, nevertheless the use of the complex $Z_0$ produces better overall accuracy in the various calculations.

At this point, a brief remark about the transmatch might be interesting. Eq 23 says that the power delivered *by* the generator (minus the transmatch loss) is equal to the power that actually goes *into* the line. This power, minus the line loss, is *delivered* to the load. If the generator is *designed* to deliver a specified power level into a certain value of resistance, for example 50 Ω, then if the line's complex input impedance, found in Eq 24, can be *transformed* to this 50 Ω by a network of some kind, then the specified power will actually be delivered, *despite* any reflections from the load. The transmatch does this job. And we note also that in this transformation, no mention is made, or needed, of the generator's internal impedance. In fact, the generator impedance does not appear in *any* of our equations or discussions. Finally, a word about the load reflections: they cause the input impedance of the line to be different from $Z_0$ (the line is an impedance transformer), and when the transmatch transforms this input impedance to 50 Ω, the effects of these load reflections are completely accounted for and the specified power is delivered.

Note that the equations that we use in Eqs 24 through 30 are exact, and do not involve the approximations that have been used in the past to simplify manual calculations. This greatly reduces the array of possible choices in problem solving. This approach has been called the brute-force approach. This is no longer true with *Mathcad* and similar programs. We are limited only by the accuracies in the various parameters, which may involve precise lab measurements that are beyond the average amateur workshop or may be found from simulations of some kind. Reliable handbook data is often used instead.

Because of the measurement inaccuracies of load impedance, matched loss, velocity factor, cable length, etc, how accurately can we calculate system performance? The equations and simulations can be used to find the *sensitivities* to these errors. Impedance values are the most difficult to know without a good measurement (or truthful simulation) method and usually represent the biggest source of errors. Also, with a calibrated transmission line sufficiently long it is possible to measure inaccessible antenna impedances using the exact equations in this article—if we are careful to eliminate errors due to common-mode currents induced by the antenna.

## Examples

To get a feel for how these equations work, the *Mathcad* worksheet of Sidebar A goes through a realistic example that we might run into in the ham shack. This example might be used as a template for a wide variety of similar exercises, including frequency sweeps, parameter sweeps and three-dimensional sweeps, that the reader can devise. In this example, meters of length, radians per meter and nepers per meter are used and are a good idea, in order to minimize confusion and errors. Matched loss, 0.57 dB per 100 feet, is converted to 0.002154 nepers per meter, a 50-foot coax length is converted to 50 × 0.3047 meters and the phase constant is in radians per meter. Coax power levels are in W, and we prefer the cable loss in dB. In this example, for a matched cable the loss would be 0.285 dB and with an input VSWR of 1.86 the loss is 0.337 dB. The additional loss is 0.052 dB. In other words, the standing waves of current and voltage produce greater losses in the copper and the dielectric, respectively. The load VSWR, if we want to know it, can be found by temporarily setting $A$, the dB per 100 feet, to 0.0. In this example the load VSWR is 1.92.

As a much more difficult example, consider at 28 MHz a 120 foot length of $50 - j2$ Ω coax with 1.0 dB per 100-foot attenuation, connected to a $200 + j100$ Ω load. Again, use meters, radians and nepers. Using Eq 24, plot $1/Z_d$ versus $d$ (distance from load) to find the first point on the coax where the real part of the admittance (the conductance) is 1/50 S. Then use the imaginary part (the susceptance) of this $1/Z_d$ to calculate the length of a shorted stub (solve Eq 25 for $d$), using the same kind of coax, that has the negative value of susceptance and therefore produces a $50 + j0$ Ω junction. Use Eq 24 again to find the coax input impedance at the generator (not *exactly* 50 Ω). Then, using Eqs 26, 28 and 29, find the actual and dB power loss in the stub, in the coax between the

## Example of SWR, Loss and Input Z for Transmission Lines

Matched Loss (dB/100 ft)

$A := 0.57$

Matched Loss (nepers/meter)

$\alpha := \frac{A}{100} \cdot \frac{3.282}{8.686}$

$\alpha = 2.154 \cdot 10^{-3}$

Length (feet)

$L_{ft} := 50$

Length (meters)

$L_m := L_{ft} \cdot 0.3047$

Velocity Factor

$VF := 0.66$

Line $R_0$

$R_0 := 50$

Load Impedance

$Z_L := 43 + j \cdot 30$

Frequency (MHz)

$f := 7.15$

Phase Constant (radians/meter)

$\beta := \frac{2 \cdot \pi \cdot f}{VF \cdot 299.8}$

Complex $Z_0$

$Z_0 := R_0 \cdot \left(1 - j \cdot \frac{\alpha}{\beta}\right)$

$Z_0 = 50 - 0.474j$

Propagation Constant per meter

$\gamma := \alpha + j \cdot \beta$

Input Impedance

$Z_{in} := Z_0 \cdot \left( \frac{Z_L + Z_0 \cdot \tanh(\gamma \cdot L_m)}{Z_0 + Z_L \cdot \tanh(\gamma \cdot L_m)} \right)$

$Z_{in} = 65.688 + 31.901j$

$|Z_{in}| = 73.025$

Assume input voltage

$E := 90.04$

Load voltage

$E_L := E \cdot \left( \cosh(\gamma \cdot L_m) - \frac{Z_0}{Z_{in}} \cdot \sinh(\gamma \cdot L_m) \right)$

$E_L = 75.292 + 15.46j$

$|E_L| = 76.863$

Power in load (W)

$P_L := E_L \cdot \overline{E_L} \cdot \mathrm{Re}\left(\frac{1}{Z_L}\right)$ $\quad P_L = 92.412$

Power input (W)

$P_{in} := E \cdot E \cdot \mathrm{Re}\left(\frac{1}{Z_{in}}\right)$ $\quad P_{in} = 99.866$

Loss (dB)

$Loss := 10 \cdot \log\left(\frac{P_{in}}{P_L}\right)$ $\quad Loss = 0.337$

Matched loss (dB)

$ML := \frac{A}{100} \cdot L_{ft}$ $\quad ML = 0.285$

Added loss (dB)

$Loss - ML = 0.052$

Input reflection coefficient

$\Gamma_{in} := \frac{Z_{in} - Z_0}{Z_{in} + Z_0}$ $\quad \Gamma_{in} = 0.197 + 0.226j$

$|\Gamma_{in}| = 0.3$

Input VSWR

$SWR_{in} := \frac{1 + |\Gamma_{in}|}{1 - |\Gamma_{in}|}$ $\quad SWR_{in} = 1.858$

stub and the load and also between the stub and the generator, assuming the appropriate value of input voltage at the generator. Then plot input impedance (preferably admittance) versus frequency to see what the transmatch (if any) requirements are. This is a very messy problem using traditional methods, but is a lot easier using *Mathcad* or similar software. Once the worksheet has been set up, it becomes part of your library. For all of this work a comfort factor with math methods and some skill with the software (easily acquired) are necessary, though. Also, a familiarity with Smith Chart graphical methods and the ARRL *Microsmith* software can add extra dimensions to our understanding.

The equations in this article apply also to parallel-line transmission lines, except Eqs 6, 8, 9, 10 and 14, that apply to coax only. The references provide more details for parallel lines.


### Acknowledgments

The problems associated with complex values of $Z_0$ with respect to the traveling voltage/current wave reflection coefficient and VSWR were realized many years ago (as shown in Notes 8 and 9). I appreciates the efforts of Dean Straw, N6BV, (at ARRL) and Frank Witt, AI1H, in bringing this to his attention, and also their comments and suggestions for this article (but any mistakes belong to me). Also, Dean's excellent TL.EXE program has been very helpful in writing this article and comparing the numbers.


Finally, the ITT handbook *Reference Data for Radio Engineers*, Transmission Lines chapter, is a gold mine of data, formulas and information (some of which can be rather intimidating) that can be used to adapt and modify the equations and procedures of this article to other applications, especially twin-lead and ladder-line, which are popular among hams. As mentioned before, many of the special-case formulas and simplified formulas are not needed by *Mathcad*. The ARRL *Antenna Book* is also highly recommended.

### Notes


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[4]Ferber, M., W1GKX, (Times Wire and Cable Co), "Coaxial Cable Attenuation," *QST*, April 1959, p 20.

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[6]Witt, Frank, AI1H, "How to Evaluate Your Antenna Tuner," Part 1, *QST*, April 1995, p 30; Part 2, May 1995, p 33.

[7]Sabin, W. E., WØIYH, "The Lumped-Element Directional Coupler," *QEX*, March 1995 p 3, and May 1995 p 8.

[8]Kurokawa, K., "Power Waves and the Scattering Matrix," *IEEE Transactions on Microwave Theory and Techniques*, March 1965, Sec IX, p 201.

[9]Johnson, W. C., *Transmission Lines and Networks*, Chapter 4, pp 93-96. McGraw-Hill, New York, 1950.

# A Packet Voice Communication System

*A simple and inexpensive system shows one way of sending voice via packet.*

By Dirench Dogruoz, Carlos Lopez, Harvind Samra, KBØPRP, and Brett Weldy

The popularity of packet-radio communication exploded toward the end of the 1980s. It was then that amateurs realized the tremendous benefits that packet communication could provide. Yet almost a full decade later, it's almost exclusively used for data and computer communication. The only other common application of packet radio is for the transfer of low-bit-rate weather satellite images. Voice and audio communication, as far as we know, has yet to be implemented with amateur packet radio.

Obviously, voice communication is common throughout the amateur world. However, a packet voice system could provide many advantages over traditional methods that use continuous data streams. These advantages include error detection and correction, packet relaying and frequency reuse. Packet relaying allows the transmission of voice to a string of other packet stations, which provides useful networking. Since the transmission of a packet uses the bandwidth for only a short duration, multiple connections can be achieved over the same frequency, or the frequencies can be reused. Therefore, a packet voice system has many features that would create an effective method of personal communication between Amateur Radio operators.

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### Background

The purpose of this article is to succinctly introduce a packet voice communication system. This system was designed and implemented as our final project for our senior-level electronics design course here at the University of Kansas. The idea of our packet voice system was suggested by our instructor, Professor Glenn E. Prescott, WBØSKX. It is his belief that this system could be easily created through the use of digital signal processing, in particular the TMS320C50 DSP Starter Kit (DSK) from Texas Instruments.

The communication system outlined offers a minimum transmission rate of 19.2 kbit/s. The design is composed of three main components: a data source and endpoint, a terminal node controller and a transceiver. Fig 1 demonstrates the basic architecture of the packet radio system.

The heart of our packet voice system is the DSK board and software from Texas Instruments. It is a simple, stand-alone application platform that allows you to experiment with the TMS320C50 digital signal processor

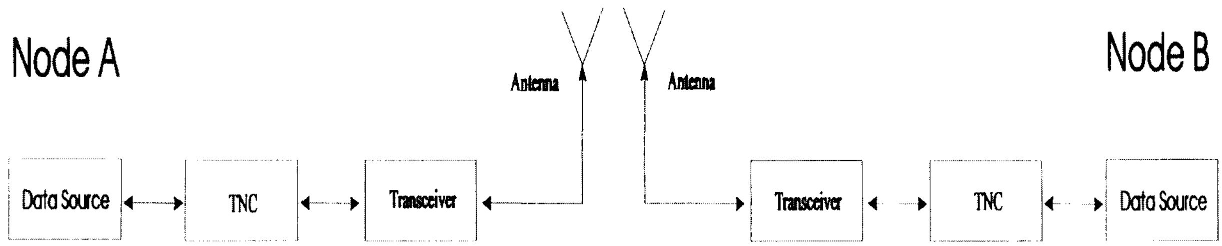


**Fig 1—Block diagram of packet voice communication system.**

to implement real-time DSP solutions. The cost of a DSK system is currently $99. It also contains both an A/D converter and a D/A converter that are connected to separate RCA audio connectors. It comes with a power supply that plugs into a standard electrical outlet. A DB-9 serial port is provided to allow a terminal or computer to communicate with the DSK board. Finally, the DSK system comes with an assembler, loader and debugger, along with some sample code for first-time users.

The data source and endpoint incorporate the DSK board. First, the data source receives an analog voice-signal input through a microphone and amplifier circuitry (needed only if the microphone's output amplitude is not large enough). The data is then sampled at 16 kHz with 14 bits per sample using the DSK, resulting in an initial data rate of 224 kbit/s. To reduce this enormous data rate, the voice samples are compressed using a continuous variable-slope delta-modulation (CVSD) algorithm to reduce the sampling rate down to 16 kbit/s. This is a significant improvement over conventional PCM coding at 64 kbit/s. The compressed voice data is then sent to a Kantronics data engine TNC that divides the continuous data stream into packets. Finally, the packetized data is transmitted using a Kantronics D4-10 transceiver. This sequence of events models the functions on the transmitter side of the system. The receiver side operates in an exactly reverse manner. The TNC receives the packetized data, depacketizes it and passes it to the receiving DSK starter kit where it is decompressed using the CVSD algorithm, converted back to analog voice and played back through a speaker.

### Configuration

The required hardware for our packet voice communication system

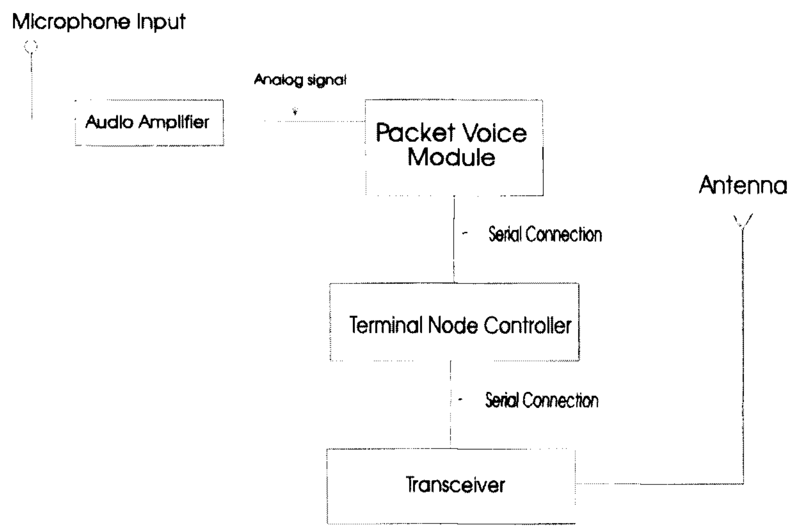


**Fig 2—Block diagram of packet voice system transmitter.**

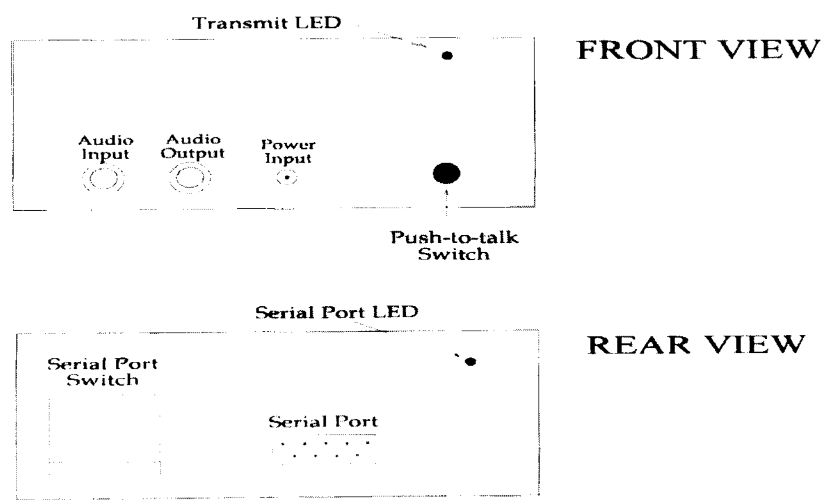


**Fig 3—Front and rear panels of packet voice module.**

consists of two packet voice modules, two specialized serial connectors, two Kantronics data engines, two D4-10 transceivers, two antennas, an audio amplifier, a microphone and a speaker. In addition, a PC is required for loading the software onto the DSK boards inside the packet voice modules.

The microphone is connected to a standard audio amplifier and the output of that amplifier is connected to the transmit input of the packet voice module. On the module is a push-to-talk switch used to send data and another switch used when loading the compression algorithm to the board. The packet voice module is connected to the Kantronics data engine via a serial cable and a specialized serial connector. The TNC is connected to the Kantronics D4-10 transceiver, and the I/O port of the D4-10 is connected to the antenna (Fig 2). The same setup is used on both sides except that on the receiver side the output of the packet voice module is connected to a speaker.

The TNC then needs to have the correct settings. The following commands indicate the necessary settings for the TNC.

cmd: MODE 19,200,NONE,8,1
cmd: INTERFACE TERMINAL
cmd: AUTOLF ON
cmd: BREAK ON
cmd: ECHO OFF
cmd: FLOW ON
cmd: FLOWR OFF
cmd: FLOWX ON
cmd: FLOWTR OFF
cmd: FLOWTX OFF
cmd: LEDS ON
cmd: RING ON
cmd: CR ON

The TNC has to be placed in conversation mode by simply typing the character *k* followed by a return.

The software is loaded into the packet voice module with a DSP loader program running on a personal computer. A null-modem serial cable connects the serial port of the personal computer to the serial port of the packet voice module. The module is programmed by typing the following DOS command with the serial port switch in the OFF position:

> dsk5l dsp

The serial port switch is then placed to the ON position. The specialized serial connector goes onto the packet voice module's serial port; the DB-9 male end goes to the module's serial port. After connecting the TNC's RS-232 port to the serial connector, the serial port switch is put back to the OFF position. Now the user is ready to send voice messages via packet.

### Operation

The packet voice communication system is very easy to use. Before the operator starts to talk into the microphone, he or she simply pushes the PTT button on the face of the packet voice module. This interrupts the program inside the DSK. The user can start to talk into the microphone once the transmit LED on the module comes on. The user has about four seconds to talk into the microphone. After about four seconds of speech the module will send the data to the TNC to be broadcast.

A couple of points should be made here. The PTT button does not need to be held down continuously. Once the button is pressed and the module's transmit LED has turned on, the button can be released. Moreover, if the button is still pressed past the four

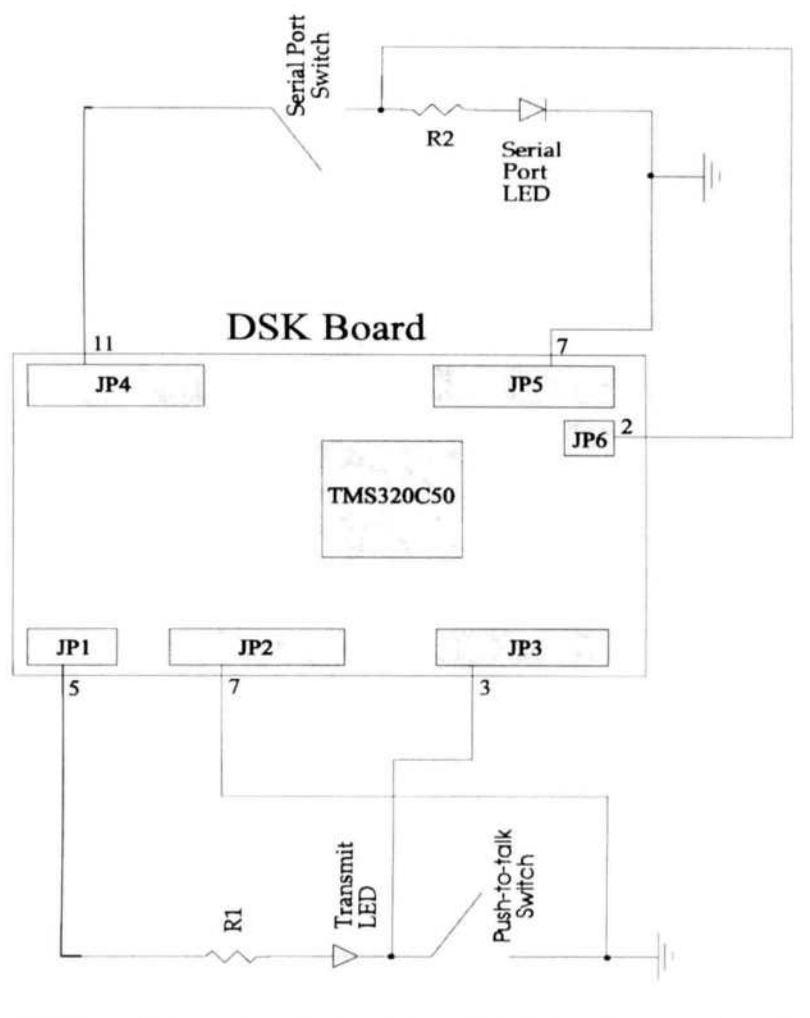


**Fig 4—Circuit diagram for interrupt switches.**


**Fig 5—Pin assignments for serial connector.**

seconds allocated for sending a message, a new message will be sampled and the previous four seconds will be discarded.

### Circuit Description

The implementation of the switches for the correct interrupt settings follows a very simple design as shown in Fig 4. The serial port switch is used to ensure that when using the serial port, the DSP board will not reset. The C50's reset signal is directly tied to pin 4 on the DSK's serial port. The serial port switch ties this pin to the +12-V supply on the DSK so that if the port is not connected, pin 4 will remain high. The PTT button is used to trigger a software interrupt on pin 3 of JP3 of the DSK board, which allows for the sampling of analog voice signals through the A/D port. Two red LEDs are used as indicators for displaying the status of pin 4 of the serial port and when the push-to-talk button is pressed.

A special serial connector is required for the TNC to properly communicate with the DSK board. Again, pin 4 on the serial DB-9 connector on the DSK can not be driven low. Unfortunately, this pin *is* driven low if the port is connected directly to the TNC. The TNC interprets this pin according to the RS-232 format; in the RS-232 protocol this is the data-terminal-ready (DTR) signal and must be asserted by whatever is connected to the TNC, in this case the DSK. To avoid this obstacle, a connector was assembled that took pin 6 from the TNC to pin 4 of the DSK. Pin 6, the data-set-ready (DSR) signal, is always high when the TNC is on. The connector also switches the transmit and receive lines, acting as a null modem. Fig 5 shows the pin assignments for the serial connector.

### Software Description

The software implemented with the DSK board has two primary purposes: (1) to compress and convert analog audio signals into digital data streams and (2) to transmit the digital data streams using the RS-232 serial protocol. Fig 6 illustrates the operation of the DSK software written for the packet voice module.

The A/D and D/A conversions are performed at a sampling frequency of 16 kHz with 14 bits for each sample. The conventional 8-kHz sampling rate for voice was not used in our system due to the limitations of the compression algorithm that was selected. The CVSD algorithm that was used ignores all frequency components above one-fourth of the sampling rate. Therefore, since a typical voice signal contains most of its energy between 300 Hz and 4 kHz, a sampling rate of $4 \times 4$ kHz = 16 kHz was chosen.

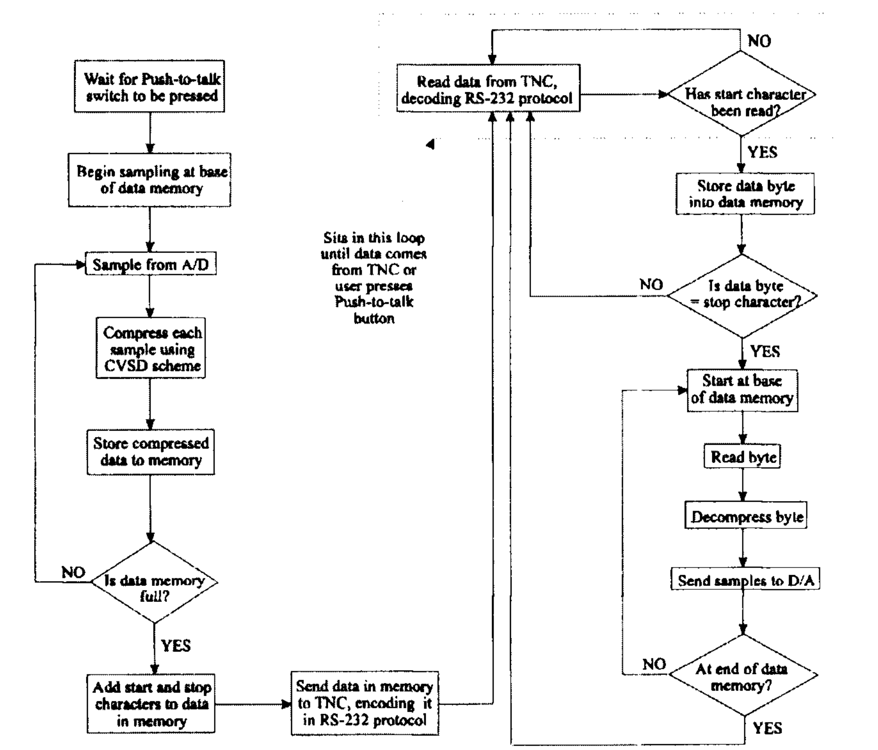


Fig 6—Flowchart of DSK software algorithm.

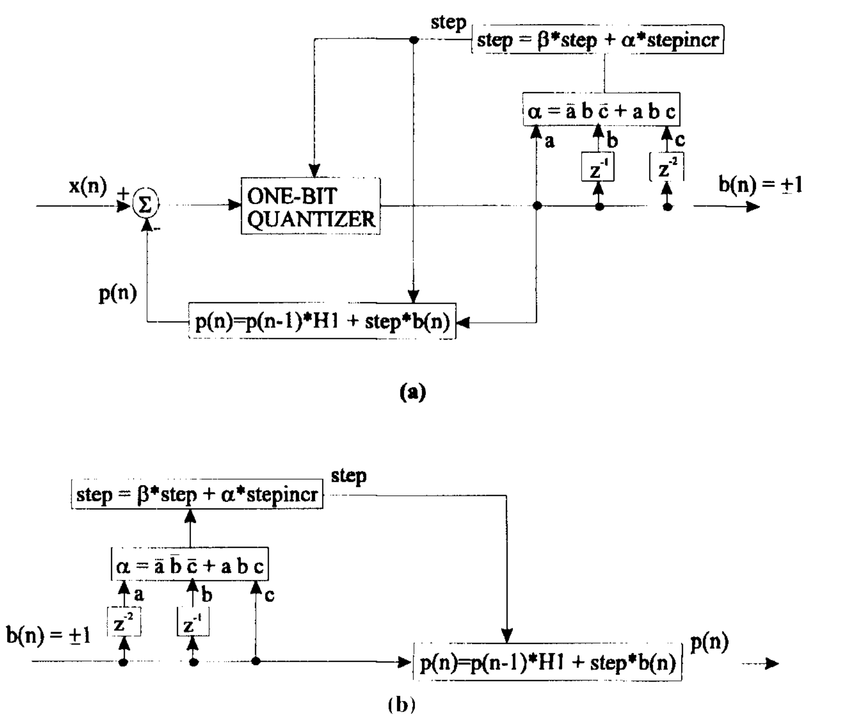


Fig 7—CVSD (a) compression and (b) decompression algorithms.

The ASCII values 1 and 254 are used for the start and stop characters, respectively. A PASS character (hex value of 16) is placed in front of characters that are used by the TNC to control communication with the terminal attached to it. The condition of the switch is monitored by connecting the switch to an external interrupt that is handled by an interrupt routine. The data memory consists of about 8000 memory locations with each location containing 16 bits or 2 bytes. Only 8 of the 16 bits of each location is used.

Fig 6 describes everything involved in the DSK software with the exception of the RS-232 formatting and CVSD compression schemes. The RS-232 encoding section of the code involves establishing a timer interrupt at 19,200 Hz and placing a start bit (0) and stop bit (1) on the ends of each data byte to be transmitted. The decoding routine is also based upon the 19,200-Hz timer interrupt. The routine waits for a start bit, combines the next 8 bits into a data byte, and detects and ignores the stop bit.

The CVSD compression and decompression algorithm implemented in the DSK code is crucial to the operation of the packet voice communications systems. It provides a compression ratio of 14, thereby greatly reducing the transmission time required for the voice message. CVSD works by attempting to measure the slope of a difference signal that is produced by taking the original signal and subtracting from it a predicted version of the original signal. Fig 7 describes the CVSD algorithm in better detail.[1]

The one-bit quantizer's quantization level is equal to the variable *step*. For the CVSD algorithm implemented in this system, $\beta$ was set to 0.995, *H1* was set to 0.9 and *stepincr* was set to 0.08. Our DSK source code can be found in the pktcvsd.zip file at the following Internet Web site: **http://www.arrl.org/files/qex/** or **ftp://ftp.arrl.org/pub/qex/**.

## Future Improvements

Clearly, our system leaves room for enhancement. First, the message limit of four seconds could be greatly increased by simply adding more RAM to the DSK board. This can be easily accomplished. Second, a better protocol could and should be implemented in terms of detecting the beginning and ending of a message. Occasionally, the DSK would receive an erroneous start or stop character. This occurred because the TNC was operating in conversation mode, which does not perform error recovery. This could be avoided by designing a more robust method of signaling and detecting start and stop characters. Third, it may be possible to handle the serial port and reset problems that we encountered and avoid using a serial port switch, specialized connectors, etc. Finally, the sound quality of the output could be enhanced by developing or implementing a more sophisticated compression algorithm. In our opinion, a digital low-pass filter on the decompressor's output could have helped.

With a simple packet voice communication system in hand, many applications could be developed. As with packet data communication, messages could be relayed and/or stored for later use. Bulletin boards could incorporate voice mail and voice messages.

**Notes**

[1]Jayant, N.S. and Noll, Peter, *Digital Coding of Waveforms*, 1984, Englewood Cliffs, New Jersey: Prentice-Hall Inc.

# Capacitance Hats for HF Mobile Antennas

*Cut-and-try works, but these calculations are an easier approach.*

By Andrew S. Griffith, W4ULD

Capacitance hats of various designs are often used with HF mobile antennas to reduce the loading inductance required to resonate the antenna and to assist in tuning the antenna to resonance. Reducing the loading inductance of a mobile antenna can significantly improve its efficiency since the loading inductance is a primary source of resistance in the antenna system. I have worked with hats on mobile antennas for years, but always on a cut-and-try basis because design information was not readily available. The recent resurgence of HF mobile operation prompted me to try to develop information that could be used to design a mobile antenna with a capacitance hat using a hand calculator. The calculation method presented in this article provides improvements in the calculation of design parameters for antennas with or without a capacity hat. Only a basic knowledge of trig functions and logarithms is required. This article does not address the calculation of feed-point impedances or matching circuits. Excellent material on these subjects may be found in the references at the end of the article.[1,2]

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## General Design Approach

If the length of a grounded vertical antenna is less than 1/4 wavelength at the operating frequency, the reactance at any point along the antenna can be estimated by considering the antenna to be a transmission line.[1,3] The portion of the antenna above the point of calculation can be considered an open-ended transmission line having a capacitive reactance equal to $-ZOW/\tan(TW)$ where $ZOW$ is the characteristic impedance in ohms and $TW$ is the electrical length of the top section or whip in degrees. For this discussion the negative sign for the top section reactance will be ignored. The bottom or base section can be considered a closed-end transmission line having an inductive reactance at its top of $ZOB \times \tan(TB)$ where $ZOB$ and $TB$ are the characteristic impedance in ohms and electrical length of the bottom section in degrees. To resonate an antenna that is less than 1/4 wavelength, a loading coil must be placed somewhere along the length of the antenna.[4] This becomes the point of calculation. At resonance, the inductive reactance of the loading coil ($XLC$)

[1]Notes appear on page 20.

plus the inductive reactance of the bottom section ($XB$) must equal the capacitive reactance of the top section ($XT$). That is:

$$XB + XLC = XT \qquad \text{Eq 1}$$

By rearranging the above equation, the loading coil reactance required for resonance is equal to the top section reactance minus the bottom section reactance. The inductive reactance of the loading coil may be converted to inductance by Eq 18 in the Appendix. This will be discussed later in an example calculation.

The loading coil itself and anything added to the antenna above the loading coil, such as a hat, will add capacitance to the top section of the antenna. This capacitance will appear in parallel with the capacitance of the top antenna element or whip and reduce the capacitive reactance of the top section. Thus the reactance of the top section will be equal to the reciprocal of the sum of the reciprocals of the capacitive reactances of the coil ($XC$), the hat ($XH$) and the whip ($XW$). The antenna reactance balance at resonance now becomes:

$$XB + XLC = \frac{1}{\frac{1}{XC} + \frac{1}{XH} + \frac{1}{XW}} \qquad \text{Eq 2}$$

If four of the reactances are known, the fifth can be calculated and converted to length, capacitance or inductance according to its source.

The equivalent velocity factor ($K$ factor) of the base section of the antenna must also be considered in the antenna design. The $K$ factor is the actual length of a resonant $^1/_4$-wavelength vertical divided by a theoretical $^1/_4$ wavelength as given in Eq 8 of the Appendix. The $K$ factor is normally in the range of 0.93 to 0.97. However, the apparent $K$ factor will increase if the antenna is mounted low on the vehicle so the feed point is close to a vertical body section. Antennas mounted on the rear deck or top of a metal body vehicle have a normal $K$ factor of about 0.95 provided the mounting surface is properly grounded to the rest of the vehicle.

## Experiments and Results

A series of experiments was run to determine the effective capacitance of some common mobile hat configurations with a variety of loading coils, bottom sections and whips. In all cases, the hats were mounted on top of the loading coil. In each experiment the resonant frequency of the antenna combination was determined with an MFJ-249 SWR analyzer.[5] The characteristic impedances of the whip ($ZOT$) and the base section ($ZOB$) were calculated from Eq 9 in the Appendix. The electrical lengths of the whip ($TW$) and the base section ($TB$) at the resonant frequency were calculated from Eq 11. The reactance of the whip ($XW$) resulted from Eq 12 and the added capacitive reactance of the loading coil ($XC$) was determined by Eq 16. The inductive reactances of the loading coil ($XLC$) and the base section ($XB$) were calculated from Eqs 14 and 13, respectively. The capacitive reactances of the experimental hats were calculated by a rearrangement of Eq 2 as shown below:

$$XH = \frac{1}{\frac{1}{XB + XLC} - \frac{1}{XC} - \frac{1}{XW}} \qquad \text{Eq 3}$$

The capacitance of the hat was subsequently calculated from Eq 19. A BASIC computer program was written to speed up the many calculations.

Since many of the experiments were run with a bumper mount, it was important to determine the $K$ factor to be applied to the base section of each antenna. This was done by measuring the resonant frequency of an 11.323-foot monopole mounted at various distances from a vertical section of the vehicle body or bumper. The $K$ factor was calculated with Eq 8. The results are shown in Table 1. The first few inches of vertical metal near the feed point are most important. For example, the data for the first line of Table 1 were obtained with the monopole mounted in place of a trailer hitch on the step bumper of a pick-up truck. The monopole was about seven inches from the tail gate of the truck but only $2^1/_4$ inches from the $6^1/_2$-inch vertical rise of the bumper. The $K$ factors of monopoles mounted on the rear deck and hood of a vehicle were normal at about 0.95. The deck and hood mounted antennas were capacitively grounded to the vehicle body with 220 square inches of aluminum foil. The experimentally determined $K$ factors were used in calculating the electrical length of only the bottom section using Eq 11.

The physical size and shape of the loading coil add both length and capacitance to the antenna. The effects could not be readily calculated or measured by assuming that the coil was a solid object. The net effect appeared to be capacitive; therefore the coil capacitance was lumped with the capacitance of the top section of the antenna. The capacitance of several loading coils was determined for several antenna configurations without hats by calculating the coil capacitance required to balance the reactances of the top and bottom sections of the antennas. The following rearrangement of Eq 2 was employed:

$$XC = \frac{1}{\frac{1}{XB + XLC} - \frac{1}{XW}} \qquad \text{Eq 4}$$

The average results are given in Table 2. These values were used in the final calculation of hat capacitances.

It was also noted that a $K$ factor of 1.0 was required for the whip to achieve consistent results. The actual

**Table 1—Effect of Vertical Vehicle Body on Antenna Length**

| *Distance from Vertical Metal Body, inches* | *K Factor* |
|---|---|
| $2^1/_4$ | 1.08 |
| $3^1/_4$ | 1.013 |
| $3^1/_2$ | 0.998 |
| 6 | 0.985 |
| 9 | 0.969 |
| 12 | 0.962 |
| 14 | 0.951 |
| 16 | 0.949 |

**Table 2—Experimental Loading Coils**

| *Length inches* | *Diameter inches* | *Turns, Wire* | *Inductance μH* | *Capacity pF* | *Construction* |
|---|---|---|---|---|---|
| $5^1/_2$ | $^7/_8$ | 106 no. 20 | 47.6 | 0.8 | Wound on $^3/_4$" CPVC |
| $5^5/_8$ | $^{15}/_{16}$ | 36 no. 12 | 5.4 | 0.6 | Wound on $^3/_4$" CPVC |
| $8^1/_2$ | $1^5/_8$ | Unknown | 182.2 | 1.35 | Commercial Mobile Coil |
| $8^1/_2$ | 3 | 69 no. 14 | 111.4 | 1.7 | Air Wound Miniductor |
| $8^1/_4$ | $2^1/_2$ | 51 no. 13 | 47.7 | 1.8 | Air Wound Miniductor |

Note: Inductance measured by dip-oscillator method. See text.

*K* factor is probably close to 0.95. The difference is no doubt compensation for some other factor that was not considered in the calculations.

The results of the work are shown in Fig 1 as a plot of hat capacitance versus hat diameter. The experimental hats are shown in the legend of Fig 1. The solid disks were constructed from circuit board. The open hats were constructed from no. 12 copper wire except that one 8-spoke wheel was constructed from 1/4-inch diameter copper tubing to confirm that wire diameters up to 1/4 inch have little effect on the reported hat capacitance. This was also confirmed by modeling some of the hats with ELNEC.[6] An 18-inch diameter wheel was also constructed from circuit board. The wheel had a 3-inch diameter hub, a 1 1/4-inch wide rim and four 1-inch wide spokes. The capacitance of this wheel fell between the solid disks and the 8-spoke wheels.

The experimental bottom sections included in the study were a group of 5/8-inch diameter masts with lengths of approximately 2.9, 5.6, 8.5 and 11.3 feet and a single 1-inch diameter mast 5.8 feet long. Whips included were 3.9 feet long by 0.09 inch average diameter, 1.79 feet long by 0.09 inch diameter and 6.9 feet long by 3/8 inch diameter. The experimental loading coils were described in Table 2.

The data in Fig 1 represented by the broken lines are actually from *ELNEC* models. In the *ELNEC* model the effects of loading coil capacitance and type of mount are not involved because the loading coil is represented by a point and the base of the antenna was mounted on a 2-foot high hill with no obstructions. Surprisingly, the experimental data fit the *ELNEC* derived lines so well that it was not worthwhile to derive new lines. The solid line representing the solid disks is plotted through the experimental data points. Most of the data fell within ±2% of the averages, and all data fell within ±10%.

Adjustable two-wire hats were also studied. The adjustable wires were mounted on top of the loading coil. Two-wire hats in the shape of a V are often used to tune an antenna to resonance with a fixed inductor and whip by adjusting the angle between the wires. The capacity of two 12-inch long no. 12 wires at several included angles is shown in Table 3. The data in Table 3 indicates that a 75-meter antenna can be adjusted over a range of approximately 170 kHz.

## Using the Correlations

The foregoing data and correlations can now be used to calculate an unknown parameter when the others are known. The parameters are the base section length and diameter, the whip length and average diameter, the inductance of the loading coil, the added capacitance of the loading coil and the type and diameter of the hat, if one is used. The added capacitance of the loading coil should be estimated by interpolating the data in Table 2. Preferably, the *K* factor of the base section should be determined by measuring the resonant frequency and length of a monopole installed on the same mount as that planned for the loaded antenna. Eq 8 should be used for this purpose. Otherwise, Table 1 should be consulted for a bumper or

**Table 3—Capacity of Two Tuning Wires**

| *Angle Between Wires, Degrees* | *Capacity pF* |
|---|---|
| 0 | 2.00 |
| 22.5 | 2.40 |
| 45 | 2.65 |
| 90 | 2.80 |

Note: No. 12 wire, each 12 inches long.

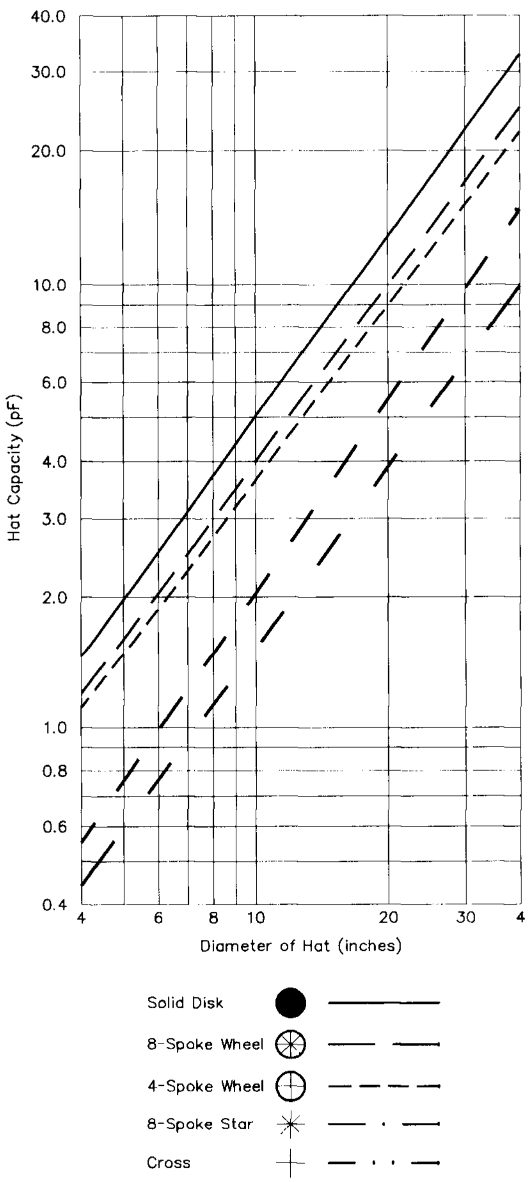


Fig 1

"trailer hitch" type mount and 0.95 should be used for an on-body mount. At 40 meters and above the effect of an error in $K$ is small since the reactance of the base mast is small compared to the reactance of the loading coil. Below 40 meters the reactance of the mast may be larger than the reactance of the loading coil. Thus any error in $K$ will be important.

In the first example, the inductance required to resonate an antenna at 7.2 MHz will be calculated. The antenna will have a 5.8 foot by 1 inch diameter base section, a 3.9 foot long by 0.09 inch average diameter whip, an 8-inch diameter 8-spoke wheel for a capacity hat and will be mounted on a bumper. It is expected that the inductance will be wound on a form that is about 0.88 inch diameter by 5½ inch long. The antenna will be bumper mounted quite close to the body of the vehicle, so a $K$ factor of 1.08 for the base section will be used. From Table 2, a loading coil of the approximate size planned will have a capacitance of approximately 0.8 pF, and from Fig 1 the hat will have a capacitance of 3.1 pF.

Step 1—From Eq 9 calculate $ZOB$ and $ZOW$ for the base section and the whip. $ZOB$ is 277.7 Ω and $ZOW$ is 398.4 Ω.

Step 2—From Eq 10 calculate the wavelength ($WL$) at 7.2 MHz. $WL$ is 136.61 feet.

Step 3—From Eq 11 calculate the electrical length of the whip ($TW$) using 1.0 for $K$. $TW$ is 10.277°.

Step 4—From Eq 12 calculate the capacitive reactance ($XW$) of the whip. $XW$ is 2197 Ω.

Step 5—From Eq 16 calculate the capacitive reactance ($XH$) of the hat. $XH$ is 7131 Ω.

Step 6—From Eq 16 calculate the capacitive reactance ($XC$) of the loading coil. $XC$ is 27,631 Ω.

Step 7—From Eq 17 calculate the summed reactance for the top section ($XT$). $XT$ is 1583 Ω.

Step 8—From Eq 11 calculate the electrical length ($TB$) of the base section using 1.08 for $K$. $TB$ is 14.152°.

Step 9—From Eq 13 calculate the inductive reactance ($XB$) at the top of the base section. $XB$ is 70.0 Ω.

Step 10—By rearranging Eq 1, the required inductive reactance is $XT-XB$ or 1583–70=1513 Ω.

Step 11—From Eq 18 calculate the inductance ($L$) required in the loading coil. $L$ is 33.4 μH.

The number of turns in the coil can now be determined from the formulas in the *ARRL Handbook*.[7]

If no hat were desired in the above example, the term $1/XH$ would be omitted from Eq 17. The required coil reactance would still be $XT-XB$. The coil reactance and inductance would be higher to resonate without the hat.

It should be apparent that the essential steps will be the same no matter which parameter is being calculated. It is a matter of which reactances are added or subtracted in Eq 2, which is the basic equation for balance of the top and bottom reactances. When you're calculating one of the top reactances it's often easier to work with Eq 2 in the form:

$$\frac{1}{XB+XLC}=\frac{1}{XC}+\frac{1}{XH}+\frac{1}{XW} \quad \text{Eq 5A}$$

To determine the reactance of a hat to bring an antenna to resonance the equation becomes:

$$\frac{1}{XH}=\frac{1}{XB+XLC}-\frac{1}{XC}-\frac{1}{XW} \quad \text{Eq 5B}$$

or

$$XH=\frac{1}{\frac{1}{XB+XLC}-\frac{1}{XC}-\frac{1}{XW}} \quad \text{Eq 5C}$$

For a second example, a 5-foot long commercial base section is available. The base section looks like a fiberglass pole. It is desired to calculate the $ZOB$ of the mast so that loading coils and hats can be designed. It is probable that the conductor inside the pole is fairly small, which means that the $ZOB$ will be high. The first step is to determine the base section reactance, $XB$. A well characterized whip and coil are available. So Eq 2 becomes:

$$XB=\frac{1}{\frac{1}{XC}+\frac{1}{XW}}-XLC \quad \text{Eq 6}$$

The term $1/XH$ is missing because the test antenna does not use a hat. The resonant frequency is determined with an MFJ-249 SWR analyzer, and the known parameters are calculated as in the steps above and plugged into Eq 6 to give a value of $XB$. The electrical length ($TB$) of the base section is calculated from Eq 11. Eq 13 is rearranged to:

$$ZOB=\frac{XB}{\tan(TB)} \quad \text{Eq 7}$$

The $ZOB$ turns out to be 409 Ω. By rearranging Eq 9 the diameter is calculated to be 0.098 inch. So it is probable that the conductor inside the fiberglass pole is either no. 12 wire or a ⅛-inch diameter rod.

When a fixed loading coil is used in an antenna design, it is important to know the inductance accurately. The best method of inductance measurement that I have found is the dip-oscillator method. The dip oscillator is used to measure the resonant frequency of the parallel circuit formed by the unknown inductance and a known value of capacitance. The dip oscillator should be coupled to a good frequency counter and measurements made with each of five to ten different silver-mica capacitors of the same value. The inductance values are then averaged. It is important to make the measurements with the minimum coupling between the dip oscillator and the coil. Minimum coupling will show the minimum perceptible dip. The *ARRL Handbook* should be consulted for further information on determining inductance by the dip oscillator method.

## Summary

A common method of designing loaded mobile antennas has treated the top whip and bottom masts as transmission lines, with the inductive reactance of the loading coil added to that of the base mast. At resonance, the reactance of the whip must equal the sum of the reactances of the loading coil and base mast. In this study, two factors have been added to improve the accuracy of the calculations. First, the added capacitance of the loading coil is lumped with the capacitance of the whip. Second, the length or $K$ factor of the base mast has been adjusted to compensate for the mounting position of the antenna on the vehicle. The capacitance of several hat configurations has been determined and a method of incorporating this capacitance in antenna design has been presented. Depending upon the accuracy of measurement of loading coil inductance, the calculations of antenna reactances can be made within 2% for most installations and within 10% for all antennas evaluated.

## About the Author

*Andrew S. Griffith, W4ULD, received a BS in chemical engineering from Virginia Polytechnical Institute in 1943 and an MS in 1947. He retired from the E.I. DuPont de Nemours Company in 1982 after 34 years in research and manufacturing. Andy became interested in ham radio as a teenager. First licensed in 1951, he held an Advanced Class license until he upgraded to Extra Class in 1983. Andy now*

## Appendix

Eq 8 $$K = \frac{L \times 4 \times F}{983.6} = \frac{L \times F}{245.9}$$

$L$=length of monopole, in feet
$F$=frequency, in MHz

Eq 9 $$ZO = 60\left[\ln\left(\frac{48 \times L}{D}\right) - 1\right]$$

$ZO$=characteristic impedance of antenna (base section or whip), in Ω
$L$=length of base section or whip, in feet
$D$=diameter of base section or whip, in inches
ln=natural logarithm

Eq 10 $$WL = \frac{983.6}{F}$$

$WL$=wavelength at the resonant frequency, in feet
$F$=frequency, in MHz

Eq 11 $$T = \frac{L \times 360}{K \times WL}$$

$T$=electrical length of base section or whip, in degrees
$L$=length of base section or whip, in feet
$K$=length factor from Eq 8 or assumed
$WL$=one wavelength at the resonant frequency, in feet (from Eq 10)

Eq 12 $$XW = \frac{ZOW}{\tan(TW)}$$

$XW$=reactance of whip, in Ω (minus sign ignored)
$ZOW$=characteristic impedance or whip, in Ω (from Eq 9)
$TW$=electrical length of whip, degrees (from Eq 11)

Eq 13 $$XB = ZOB \times \tan(TB)$$

$ZOB$=characteristic impedance of base section, in Ω (from Eq 9)
$TB$=electrical length of base section, in Ω (from Eq 11)

Eq 14 $$XL = 2 \times \pi \times F \times L$$

$XL$=inductive reactance of loading coil, in Ω
$L$=inductance of loaing coil, in μH
$F$=frequency, in MHz
π=3.141593

Eq 15 $$XBT = XB + XL$$

$XBT$=total inductive reactance of base section, in Ω
$XB$=reactance of base section, in Ω (from Eq 13)
$XL$=reactance of loading coail, in Ω (from Eq 14)

Eq 16 $$X = \frac{10^6}{2 \times \pi \times F \times C}$$

$X$=capacitive reactance of hat or loading coil, in Ω
$F$=frequency, in MHz
$C$=capacity of hat or coil, in pF (from Fig 1 or Table 2)
π=3.141593

Eq 17 $$XT = \frac{1}{\frac{1}{XW} + \frac{1}{XH} + \frac{1}{XC}}$$

$XT$=total reactance of top section, in Ω
$XW$=reactance of whip, in Ω (from Eq 12)
$XH$=reactance of capacitance hat, in Ω (from Eq 16)
$XC$=reactance of loading coil capacity, in Ω (from Eq 16)

Eq 18 $$L = \frac{XLC}{2 \times \pi \times F}$$

$L$=inductance of loading coil, in μH
$XLC$=inductive reactance of loading coil, in Ω (from Eq 14)
$F$=frequency, in MHz
π=3.141593

Eq 19 $$C = \frac{10^6}{2 \times \pi \times F \times X}$$

$C$=capacitance of hat or loading coil, in pF
$F$=frequency, in MHz
$X$=reactance of hat ($XH$) or coil ($XC$), in Ω
π=3.141593

*participates in the VE program and is primarily interested in the technical aspects of ham radio.*

**Notes**

[1]*ARRL Antenna Book*, 17th Edition 1994, Chapter 16.

[2]Byron, B., W7DHD, "Short Vertical Antennas for the Low Bands," *Ham Radio*, Part 1, May 1983, pp 36-40; Part 2, Jun 1983, pp 17-20.

[3]Michaels, Charles J., W7XC, "Evolution of the Short Top-Loaded Vertical," *QST*, Mar 1990, pp 26-30.

[4]Capacity loading may be used also, but the size of a hat without an inductance will usually be too large for a mobile antenna.

[5]MFJ Enterprises, Inc, Box 494, Mississippi State, MS 39762.

[6]*ELNEC* Antenna Software, Roy Lewallen, PO Box 6658, Beaverton, OR 97007.

[7]*1996 ARRL Handbook for the Radio Amateur*, ARRL.

# A Better and Simpler A/D for the DDC-Based Receiver

*Yet another improved front end brings the DDC-based receiver closer to superhet performance*

By Peter Traneus Anderson, KC1HR

Burr-Brown has just introduced a 12-bit, 25-MSPS (megasamples per second) A/D converter, the ADS801U.[1] The ADS801U permits building a very simple improved front end for my digital downconverter- (DDC) based receiver.[2,3,4,5] The receiver's intermodulation-distortion dynamic range (IMDDR) is limited by the ADS801U, and the blocking dynamic range (BDR) is limited by the noise floor of the DDC.

The dynamic linearity specs of the ADS801U are not as good as those of the Analog Devices AD9042, but the ADS801U costs $50, versus $250 for the AD9042.[6] The AD9042 needs no active gain ahead of it, just the passive voltage gain of an impedance transformation. However, it is difficult to protect the $250 AD9042 from transients without degrading its linearity.

Fig 1 shows the circuitry. The double-tuned preselector is still built using a pair of molded RF chokes close together for magnetic coupling and a dual 365-pF variable capacitor. Choose the inductance to resonate in the desired band. I use plug-in coils—a band switch can also be used—and 150-µH chokes to cover most of the AM broadcast band and the 160-m band, and 6.8-µH chokes to cover the 80-, 40- and 30-m bands. 1.5-µH chokes are used to cover the 20- and 15-m bands. Smaller chokes would be better for these bands. These bands are between one half and one times the clock frequency of 25 MHz, so they are aliased in the A/D; the 20-m band shows up from 25–14=11 MHz to 25–14.35 =10.65 MHz, and the 15-m band shows up from 25–21=4 MHz to 25–21.45=3.55 MHz.

The 20-m and 15-m sidebands are swapped by the aliasing, so the usual 20-m and 15-m USB signals become LSB signals to the DDC. The usual signals on the lower bands are already LSB, so the DDC needs to be programmed only for LSB signals.

The ADS801U makes an especially simple front end for a receiver as its reference is internal, and an LM733 video amplifier can drive the ADS801U analog input. The LM733 has a differential output and the ADS801U has a differential input, so they are a good match.

The LM733 has enough differential

[1]Notes appear on page 24.

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output swing to drive the ADS801U to full-scale input (4-V peak-to-peak) without clipping and cannot generate enough output swing to damage the ADS801U. Running the LM733 from 6- or 7-V supplies would increase the linear swing of the LM733 at the expense of added power supplies.

The LM733 pin numbers shown are those of the 14-pin DIP package. LM733 is National Semiconductor's name for the part. Another vendor's 733 video amplifier will work as well. The 733 was one of the original linear ICs introduced by Fairchild in the late 60s and early 70s, and its gain of 100 at a bandwidth of 90 MHz is impressive even today. The 592 video amplifier is the same, except the minimum gain (all gain pins open) is zero rather than 10. The 592 lacks the 733's internal resistor between the gain pins.

The 733 is run at its minimum gain of 10 (for differential outputs), by leaving the gain pins open. Fig 1 shows a resistor between the gain pins to show where a resistor may be added to increase the gain of the 733 (or provide any gain in the case of the 592). Lower resistor values provide more gain at the expense of poorer linearity.

If you use the 592, you can place a series-tuned circuit between the gain pins to get a band-pass preamp. The 733 has a second function: it protects the more-expensive A/D from transients.

At my location I receive two strong AM broadcast stations at 1230 and 1390 kHz. By peaking the preselector between these two frequencies I was able to set the signal levels of the two stations to be equal, with each 14 dB down from full scale in the A/D.

By tuning the DDC to various intermodulation (IMD) products and measuring the audio signal level, I was able to estimate the IMD products. I found the second-order products were 66 dB down from each signal alone and the third-order products were at least 84 dB down. The noise floor of the DDC limited the third-order measurement.

For comparison, the ADS801U spec sheet shows (for input signals at 4.4 and 4.5 MHz) second-order products 78 dB down and third-order products 72 dB down. The distortion of the LM733 is unspecified, but should be low at the minimum gain, given the internal circuit design and wide bandwidth (120 MHz at the gain of 10 used here) of the LM733.

The ADS801U full-scale input voltage is $4\text{-}V_{pp}=2\text{-}V_{pk}=1.4\text{-}V_{rms}$. The LM733 has a differential gain of 10, so the full-scale input voltage at the

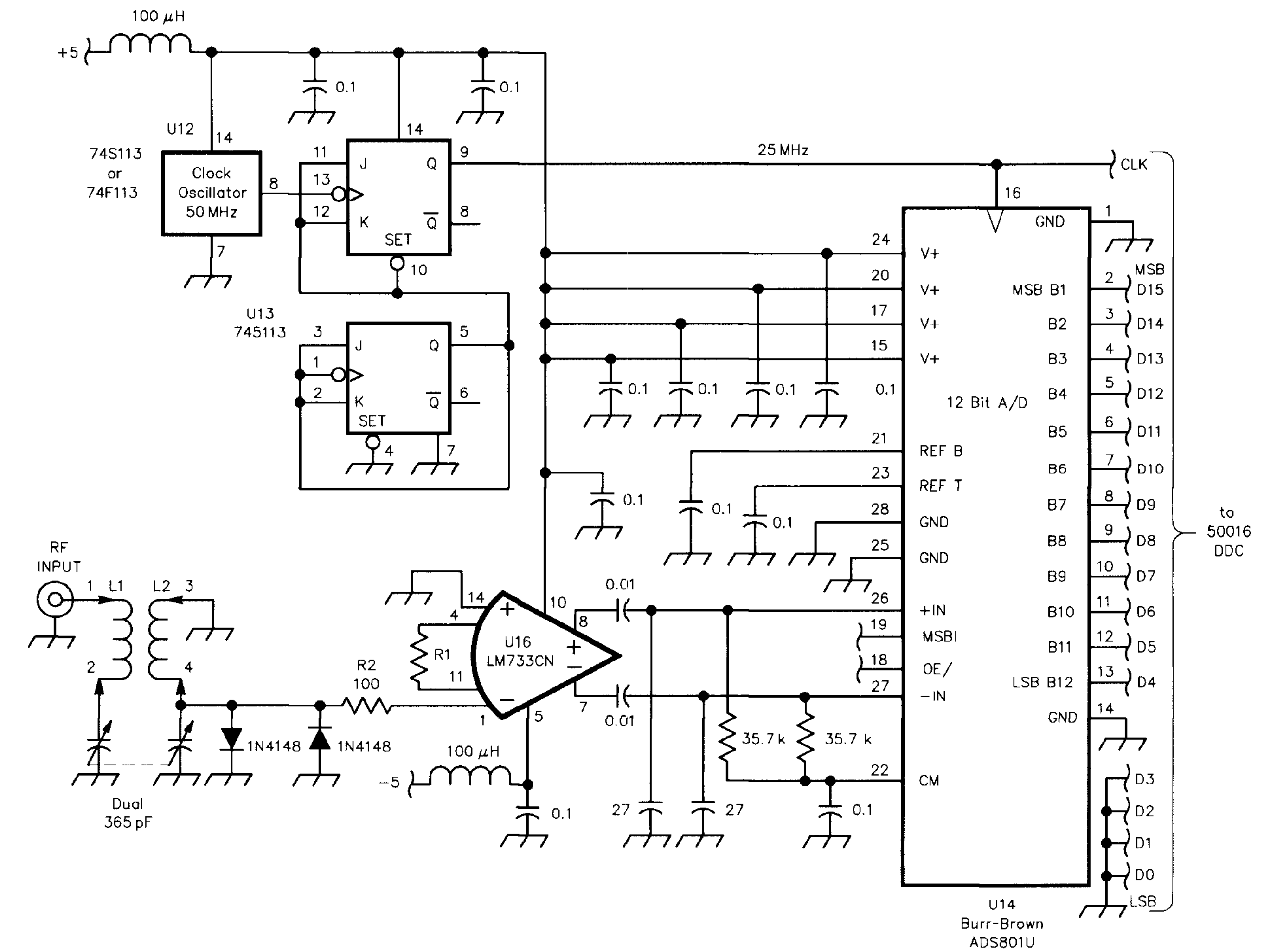


Fig 1—Preselector, preamp and 12-bit A/D converter for the DDC-based receiver. Gain-setting resistor R1 is omitted for the minimum-gain condition discussed in the text. Lower values of R1 give higher RF gains at the cost of reduced linearity.

LM733 input is 140 $mV_{rms}$. This provides excellent large-signal-handling capability.

One count in the ADS801U corresponds to 1/4096 of full scale, so one count is 35 $mV_{rms}$ at the LM733 input. The wide-band noise of the LM733 input is rated to be about the same, so the quantization noise of the ADS801U is swamped by the analog noise of the LM733, even with the LM733 set to minimum gain.

Connecting the LM733 clearly increases the activity on the lowest bits of the ADS801U (even with the LM733 input grounded), so the LM733 does indeed control the noise level.

The ADS801U is typically rated for a 66-dB wide-band signal-to-noise ratio (SNR). The effective noise bandwidth is one half the sample rate, or 12.5 MHz. If the DDC is set for a bandwith of 2000 Hz (used for SSB reception), the ratio of input bandwidth to output bandwidth is 6250.

This bandwidth reduction has a noise-averaging effect, often called process gain. For random-noise signals, the noise power is reduced by the bandwidth ratio. Equivalently, the noise voltage is reduced by the square root of the bandwidth ratio, in this case 79=38 dB.

The DDC gain is set so that a full-scale input CW carrier produces a full-scale output CW carrier. Thus, the DDC signal gain is unity. The process gain increases the SNR by 38 dB, to give an expected SNR of 104 dB.

When I listened to the receiver's noise floor, the "noise" had a tonal quality that true noise shouldn't have. Checking with an oscilloscope at the D/A output showed a low-level tone in the center of the audio passband, at one fourth of the audio sample rate.

This tone is the classic Weaver center-of-the-passband spur (called F_weaver in Fig 2) and is seen digitally as a 1-count peak-to-peak square wave when the DDC output is scaled to 17-bits peak-to-peak full scale.[7] In an analog receiver, the spur is due to dc offsets in the mixers or low-pass filters. Here, the spur is due to round-off or truncation errors in the DDC. At 6 dB per bit, the spur is 102 dB down from full scale.

The lesson here is that the DDC has mathematical errors that limit the DDC output SNR to the Harris-specified 102 dB.[8] Thus, the DDC noise floor is higher than the analog noise floor! Reducing the bandwidth does not reduce the DDC noise floor, so narrowband operation is even more DDC-limited.

Connecting an antenna causes an increase in noise in the static-laden lower bands. For higher bands, the LM733 gain can be increased by connecting a resistor across the gain pins, at the expense of poorer linearity and poorer large-signal-handling characteristics.

How good must a DDC be for a really good receiver? Say the full-scale input is 1 $V_{rms}$=13 dBm. Say the receiver is thermal-noise limited and has a minimum passband of 100 Hz. Then the noise level is 1 nV/sqr(Hz)×sq (100 Hz)= 10 $nV_{rms}$=13 dBm–160 dB= –147 dBm.

The dynamic range at the DDC output is then 1 V / 10 nV= 100,000,000=160 dB. At 6 dB per bit of resolution, this is 160/6 = 27 bits.

For a margin of safety and to use a standard word size, say we calculate to 32 bits of accuracy at the DDC output. This is twice the bits of accuracy in the Harris 50016 and requires roughly four times as many gates. However, monolithic gates are very cheap and processes have gotten denser since the already-small 50016 chip was designed. Thus, a 32-bit DDC is thoroughly practical once there is a demand for it.

Here is a challenge for receiver designers: Develop a receiver which overloads at 1 $V_{rms}$ or higher, has a noise floor equal to the thermal noise of a 50-Ω resistor at room temperature and has no detectable spurs whatsoever (all spurs are masked by the thermal noise).

This requires a 32-bit DDC and a *very* good RF digitizer. The digitizer must have an A/D that is thermal-noise limited, has 1 $V_{rms}$ full-scale range and is linear to 27 bits.

Note that the A/D output does not have to be 27 bits wide to meet these specs. In good RF A/Ds, the linearity is better than the word length. The time base for the A/D must have a phase

| | | | | |
|---|---|---|---|---|
| F_weaver = f"/4 | = | 1750 | 350 | Hz |
| FIR_decimation = f'/f" | = | 2 | 2 | |
| desired f" = 4*F_weaver | = | 7000 | 1400 | Hz |
| desired f' = 2*f' | = | 14000 | 2800 | Hz |
| -3dB_bandwidth = 0.14*f' | = | 1960 | 392 | Hz |
| audio lo band edge = F_weaver - 0.07*f' | = | 770 | 154 | Hz |
| audio high band edge = F_weaver + 0.07*f' | = | 2730 | 546 | Hz |
| -102dB_bandwidth = 0.2*f' | = | 2800 | 560 | Hz |
| F_clk = fs | = | 25000000 | 25000000 | Hz |
| HDF_decimation_ratio = fs/f' = R | = | 1786 | 8929 | |
| f'_exact | = | 13997.8 | 2800.0 | Hz |
| f"_exact | = | 6998.9 | 1400.0 | Hz |
| f_weavor_exact | = | 1749.7 | 350.0 | Hz |
| Ceiling[5*log_2(R)] | = | 55 | 66 | |
| Shift = 75-Ceiling[5*log_2(R)] | = | 20=14hex | 9 | |
| Scale_factor =2^Ceiling[5*log_2(R)]/R^5 | = | 1.98263374 | 1.300066594 | |
| 2^16*Scale_factor | = | 129934=1FB8Eh | 85201=14CD1h | |
| f_IQCLK = 100 * (top of passband) | = | 274725 | 54705 | Hz |
| f_IQCLK/f" | | 40 | 39 | |

**Fig 2—Calculations of various parameters used to set up the 50016 DDC in my previous articles (most symbols are from the Harris HSP50016 data sheet). f_IQCLK is used both to clock serial data into an audio D/A converter and to clock a Maxim switched-capacitor analog antialias filter following the D/A converter. The Maxim filter has a cut-off frequency equal to 1/100 of its clock frequency. The "ceiling" of a number is the lowest integer larger than the number. For example, the ceiling of 3.1416 is 4.**

noise low enough to support this level of linearity.

A few people have asked my how I calculated various parameters to be programmed into the 50016 DDC. Fig 2 shows these calculations. Refer to my previous articles for the control words I write to the DDC and to the 50016 spec sheet for explanations of the meaning of the parameters.

This is the third A/D I've used in my receiver. It's the simplest and best-performing so far. Better A/Ds exist, but the Harris 50016 DDC is becoming the limiting factor. In that sense, the present design is optimal for me for now. Once really good A/Ds are available, 32-bit DDCs will be needed.

**About the author**

*A radio amateur since 1980, Peter Traneus Anderson holds an Advanced Class license and has been learning how to build radios for 40 years. He received an AB in physics from Wesleyan University in 1970 and an MA in physics from Dartmouth College. He is currently studying for a PhD in electrical engineering at the University of Vermont and has been employed since 1980 at Polhemus Incorporated, where he is involved in the development of sensing systems operating in the 30,000-meter band. His areas of technical interest in Amateur Radio include the development of high dynamic range RF A/D converters and the use of these converters to build all-digital HF receivers.*

**Notes**

[1]ADS801U data sheet, Burr-Brown, PO Box 11400, Tucson, AZ 85734, 520-746-1111. For immediate product information, call 800-548-6132.

[2]Anderson, P. T., "A Simple SSB Receiver Using a Digital Down Converter," *QEX*, Mar 1994, pp 3-7.

[3]Anderson, P. T., "A Better A/D and Software for the DDC-Based Receiver," *QEX*, Nov 1994, pp 11-15.

[4]Anderson, P. T., "A Simple CW Demodulator for the DDC-Based Receiver," *QEX*, Feb 1995, pp 6-10.

[5]Anderson, P. T., "A Simple CW Transmit VFO for the DDC-Based Receiver," *QEX*, Jan 1996, pp 20-25.

[6]AD9042 data sheet, Analog Devices, One Technology Way, PO Box 9106, Norwood, MA 02062-9106, 617-329-4700.

[7]Anderson, P. T., "A Different Weave of SSB Receiver," *QEX*, Sep 1993, pp 3-7.

[8]HSP50016 data sheet, Harris Semiconductor, 1301 Woody Burke Road, Melbourne, FL 32902, 407-724-3000. You can get the data sheet from Harris's AnswerFAX 407-724-7800 or from the Internet at **http://www.semi.harris.com**. (Harris also has some nice cheap, fast 10-bit A/D converters.)

# *Conference Proceedings Available*

**Central States VHF Society Conference**

The 30th Annual Conference of the Central States VHF Society was held July 26-28 at the Thunderbird Hotel & Convention Center in Bloomington, Minesota. Here is a summary of papers presented. Conference Proceedings are available from ARRL. ISBN: 0-87259-563-3; cost is $12, plus shipping; order number: 6533.

A Brief History of the Central States VHF Society

Chambers Award Recipients

Wilson Award Recipients

Microstrip "Patch" Antenna Design; Roger A. Cox, WBØDGF

VHF/UHF Log-Periodic Antenna Design; Roger A. Cox, WBØDGF

Stochastic Resonance; Lawrence E. Stoskopf, NØUU

Moving VHF Into the Mainstream: *CQ VHF* and You; Rich Moseson, NW2L

Multiple Hop Sporadic E Path Loss; Kenneth L. Boston, WF9X

Tropospheric Propagation Observations; Mel Larson, KCØP

Comparative Analysis on the Effects of Dish Antennae Size and Solar Flux on Measured Sun Noise; Melvin B. Graves, WRØI

Phase 3D, A New Era for Amateur Satellites; William A. Tynan, W3XO, and the Phase 3D Design Team

Project *Argus* and the Challenge of Real-Time All-Sky SETI; H. Paul Shuch, N6TX

GaAs FET Pre Amp Cookbook #3; Kent Britain, WA5VJB

EMR, Health and Weak Signal DXing: New Rules and New Research; Wayne Overbeck, N6NB

Communicating With Light; Bryan Ward, N5QGH

VHF Contesting in Europe; Keith Naylor, G4FUF

Moonbounce Comes of Age; Barry Malowanchuk, VE4MA

Microwave EME Communication; Al Ward, WB5LUA

Using the Chaparral 11 GHz Superfeed at 10.368 GHz; Paul Wade, N1BWT

"Reach for Space": Flight 9; Jerome Doerrie, K5IS

Competitive Multi-Op Mountaintopping; Tim Marek, NC7K

Restoring Passband Tuning to the ICOM IC-X75 Series Radios; Dave Phillips, W7GZ

Make Your HTX-100 a Flexible IF Transceiver; Rus Healy, NJ2L

ARRL January VHF Sweepstakes Grid Records; Ken Ramirez, KP4XS/W4

ARRL June VHF QSO Party Grid Records; Ken Ramirez, KP4XS/W4

ARRL September VHF QSO Party Grid Records; Ken Ramirez, KP4XS/W4

A Cool 1500 Watt Amplifier for 432 MHz; Steve Gross, N4PZ

Preliminary Tower Analysis Techniques; Steven H. Sawyers

An Andrew Cable Primer; Tom Whitted, WA8WZG

A Comprehensive Bibliography of UHF & Microwave Articles; Paul Husby, WØUC

CSVHF 1994 Antenna Measurement Results

CSVHF 1994 Preamplifier Measurement Results

Central States VHF Society 1995 Antenna Gain Results

Central States VHF Society 1995 Noise Figure Results

North American VHF and Above DX Records

**Eastern VHF/UHF Conference**

The 22nd Annual Eastern VHF/UHF Conference was held August 23-25, 1996, at the Quality Inn and Conference Center, Vernon, Connecticut. Here is a summary of papers presented. Conference Proceedings are available from ARRL. ISBN: 0-87259-567-6; cost is $12, plus shipping; order number: 5676.

The Eastern VHF/UHF Conference Experience; Stan Hilinski, KA1ZE, Chairperson

The Eastern VHF/UHF Conference Trivia Quiz; Stan Hilinski, KA1ZE, Chairperson

*Antennas*

Application of Circular Waveguide with an 11-GHz TVRO Feed; Bruce Wood, N2LIV

East and West Combine on 10 GHz/ Part I - (3-95) Development; Bruce Wood, N2LIV, Dick Knadle, K2RIW, Ron Videtto, N2NKJ

East and West Combine on 10 GHz/ Part II - (7-96) - Results; Bruce Wood, N2LIV, Dick Knadle, K2RIW, & Ron Videtto, N2NKJ

*Equipment Design & Modification*

An Image-Phasing Transverter for 10.368 GHz; Doug McGarrett, WA2SAY

Comparison of TVRO LNB's on 10 GHz; Bruce Wood, N2LIV

HF/VHF "All-Band" TVI Filter; Dale Clement, AF1T

Improving the Vibroplex Brass Racer Iambic Paddle; Chris Fagas, WB2VVV

Kenwood TS-450S & TS-690S Low Power Modifications; Bruce Wood, N2LIV

Milling with a Drill Press; Ken Schofield, W1RIL

Mini-Circuits MMIC Component Values - Chart; Chris Fagas, WB2VVV

Noise: Measurement and Generation; Paul Wade, N1BWT

Sizing Concrete Piers for Tower Legs; C.R. MacCluer, W8MQW & W.E. Saul, P.E.

TELETEC Solid State VHF & UHF Amplifiers, Product Review; Fred Stefanik, N1DPM

Tweaking the AGC Time Constant of the Kenwood TS-700A; Chris Fagas, WB2VVV

Using the Chaparral 11 GHz Superfeed at 10.368 GHz; Paul Wade, N1BWT

Svetlana 4CX400A Improves 8930 VHF Amplifiers; Ron Klimas, WZ1V

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An Introduction to VHF/UHF Weak-Signal Operating; Jeff Klein, WA2TEO & Fred Stefanik, N1DPM

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50 MHz Beacons; Compiled by Martin Harrison, G3USF

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*Testing*

Antenna Gain Measurements, Twenty - First Eastern VHF/UHF Conference

Noise Figure Measurements, Twenty - First Eastern VHF/UHF Conference

*Transmission Lines*

Common Coaxial Cable Characteristics - Chart; Chris Fagas, WB2VVV

Typical 10 GHz Transmission Line Insertion Loss; Chris Fagas, WB2VVV

*Reprints fom QEX Microwave Experimenters' Sampler*

More on Parabolic Dish Antennas (December 1995 *QEX*); Paul Wade, N1BWT

A "Fool-Resistant" Sequenced Controller and IF Switch for Microwave Transverters, (May 1996 *QEX*); Paul Wade, N1BWT