winding your own single ended output transformer part one

by Jim Flowers

I've read well written articles about building single-ended tube amplifiers which made me want to make one for myself. The reality of completing such a project always ran into the same snag, I'm too cheap to spend the big dollars for the output iron. Suitable power supply transformers can be found as surplus, but I apparently live in an output transformer poor part of the country. I'm also amazed by what some of you can find in your junk box or at the swap meet. Well maybe not you, but some of the guys who author the tube amplifier articles in Sound Practices or Glass Audio can go to their junk box and unearth a spare Acrosound transformer for the type 50 output tubes he has collecting dust on a shelf.

I thought I might some day cure my case of junk box envy by winding my own single-ended output transformer (SE OT). A decision to build the Superwhamodynes cemented the commitment to build the OT at the same time. The information I used for the design was gleaned from many sources found in several libraries. (See Table 1 for a reference list.) I never found a "How To Build A Single-Ended Output Transformer From Scratch" cookbook. I hope to present what I did in such a way that the reader can not only follow along, but leverage the information for his own design as well.

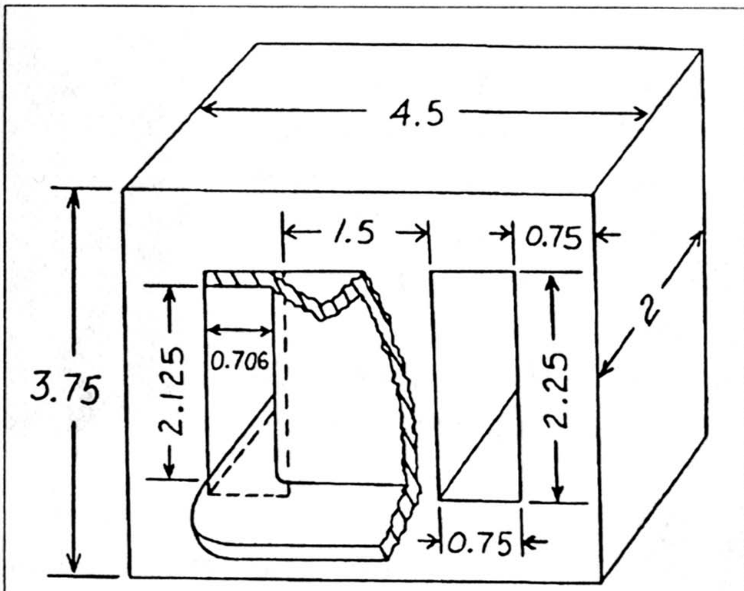

figure 1 - scrap transformer

A Caveat

Successfully building an audio transformer will emphasize the difference between design and implementation of that design. This is mainly because the integrity of the mechanical structure contributes to the electrical response in a way that is not adequately accounted for by the design formula. Even if the equations and models were complete, the builder's craftsmanship and attention to detail would still factor heavily into the success of turning schematic into product. In other words, the math will only get you so far, and at some point, you just have to build the darn thing and see how it turns out. In this case, the cost of learning by mistake is mostly one of time not money. And it can be a lot of time too. Carefully winding over a half mile of wire inch by inch is better measured with a calendar than a stopwatch.

Some details are left out, but the references should help fill in the blanks. These references can help turn black art into science. Certainly you have friends who shake their heads in disbelief at your pursuit of vintage electronics, and winding you own transformer will only further convince them of your lunacy. But besides that kind of fun, you'll come away with a better understanding of how a transformer is made; and consequently, a better appreciation of why a good SE OT costs as much as it does (and superb one even more than that).

Core Salvage

I thought a 10 watt design would be a good starting point and set out to find suitable materials. The basic ingredients of an OT are laminations, wire, and insulation. A pair of scrap $5 power supply transformers were chosen to provide the laminations. Figure 1 shows the general dimensions of what I got for my five dollars.

A transformer winding shop would probably be willing to sell laminations (and wire and insulation too). If you prefer to cannibalize a surplus transformer as I did, I suggest the following rule of thumb when choosing a victim: allow at least one pound of weight for every watt of output power. Pick a transformer with thin laminations that you think you can get apart - avoid encapsulated or canned iron.

Clean up the outside, strip the paint, and remove any mounting hardware if possible. The transformers I disassembled had a potting compound (looked similar to epoxy) that held the screws in so tight that I broke a screw trying to remove it. Heating the transformer in an oven (about 300 degrees) softened the,potting compound so that the screws came out easily.

Once the screws are out, the core can be disassembled. I have destroyed the first lamination of every transformer that I had to pry apart. This outer most lamination was made of thinner material and shaped differently from the others (an elongated E). I think this is called a keeper. Whatever its real name was, it's called trash now.

To remove a lamination, I cut it free from the core using a razor blade knife and a putty knife. The I-shaped laminations come free easily using only the razor knife. The E-shaped laminations are a little trickier because the middle section (tongue) is buried inside the coil bobbin. Work the razor knife around the outside edge and then slide the putty knife into the center of the core. The lamination should pop free without prying or any permanent bending.

The core required frequent trips to the oven to keep the potting compound soft enough to pull apart. That's not so bad as it provided worthwhile rests from a rather tedious job. You'll need to experiment to find the best temperature to work with. I used a Black & Decker Workmate to hold the hot transformer steady while I pulled it apart.

After removal, each lamination may retain a splotchy, thin skinned covering of potting residue on its surface. While this can be removed with solvent, I left it on to use as a "glue" when the core is reassembled. Clean off any large accumulations than might keep the laminations from stacking neatly.

After the core is apart, the coil can be unwrapped. Salvage the terminals for later use. Unwrapping the coil reveals the inner construction; this might provide useful tips on how to assemble the winding later. I didn't salvage the wire; it was either too short or the wrong gauge. I prefer to wind with straighter, less kinked wire anyway. (Winding is challenging enough with fresh wire.) The thick wire in your scrap transformer might be of more use. Perhaps as a 0.36 mH air core inductor for a loudspeaker crossover.

If you are not interested in coil archeology, a hacksaw might be the best way to get at the core. Use the finest pitch blade available to make several cuts on both sides. After cutting, split the winding and peel it off of the core. This procedure can also be helpful when modifying (as opposed to cannibalizing) a power supply transformer. Normally, the primary winding is next to the core and it's only the secondary windings that are to be removed. Surgical precision using the, hacksaw can remove the secondary windings while leaving the primary intact.

Take care not to damage the coil bobbin. I suggest that you don't take apart the transformer until your design is finished and you are ready to build.

Electrical Design

I figured an 845 could live a long and relatively stress free life when developing only 10 watts of output power. Furthermore, there's plenty of room for more power should I decide I need it. But for the moment, the operating conditions are approximately 700 volts on the plate, -86 volts on the grid, and 90 mA bias current.

According to RCA, the 845's plate resistance measures 1700 ohms.One rule of thumb estimates loading a triode at 4 times its plate resistance. Another rule says that loading a triode at 2 times its plate resistance gives maximum power output. Let's choose somewhere in between, say 3 times the plate resistance. Therefore, the OT needs to make an 8 ohm speaker load appear to be 3 x 1700 = 5100 ohms. The required OT turns ratio N is:

N = sqrt(5100 / 8) = 25.3

For every 25 turns of wire on the primary, there is 1 turn of wire on the secondary. The next step is to determine the total number of turns needed for the primary.

For those of you uncomfortable with using RCA's specification of a 1700 ohm plate resistance, or the determination of a suitable load, forget about the output tube for now and just think of this as a design for a single-ended 5.1 kohm transformer at 90 mA of bias current.

The primary winding of the OT is an inductor. When the OT is coupled to the output tube, a high-pass series RL circuit is formed. Ignoring the winding resistances, R is the parallel combination of the output tube plate resistance and the "transformer impedance"; i.e., the load resistance reflected into the primary (5100 ohms nominal). For a series RL circuit, the -1dB point occurs approximately where the inductive reactance equals twice the series resistance:

wL = 2R = 2 x pi x (Frequency at -1dB) x L

In this case, the frequency cutoff is inversely proportional to inductance. More inductance means a lower frequency cutoff point. As above, for the 845 use Rp = 1700. Determine the minimum inductance required for 1dB down at 20 Hz. Rearranging and solving for L:

L = 2R / [2 x pi x (Frequency at -1dB)] =
(1700 // 5100) /(pi x 20) = 20.3 H

Note that this is a minimum inductance figure. Also note that this is more stringent than the typical specification of -3dB. The -3dB point requires half this much inductance (10 H). Phase shift is about 27 degrees at -1dB; whereas, it's 45 degrees at -3dB. According to the Radiotron Designer's Handbook, choosing the -1dB point results in fairly low distortion, but still lower distortion ("good fidelity" as they put it) requires keeping the phase angle less than 15 degrees. To get only 15 degrees, the response must down only 0.3dB at 20 Hz, and that requires 1.9 times as much inductance as calculated above.

There's another reason for increasing the inductance. The harmonic distortion is directly proportional to the ratio of the primary reactance to the source resistance (see references 11, 12). The distortion is lowered as the ratio climbs (at this point the ratio = wL/R = 2). Improving the ratio requires lowering the resistance or increasing the inductance. The resistance can be lowered by changing to an output tube whose plate resistance is smaller. Since the tube has already been specified, the remaining option is to raise the primary inductance.

Based on the hope of phase and distortion improvements, I chose to bump the design goal an unscientific 50% higher. That makes the design value approximately 31 H. You may decide otherwise.

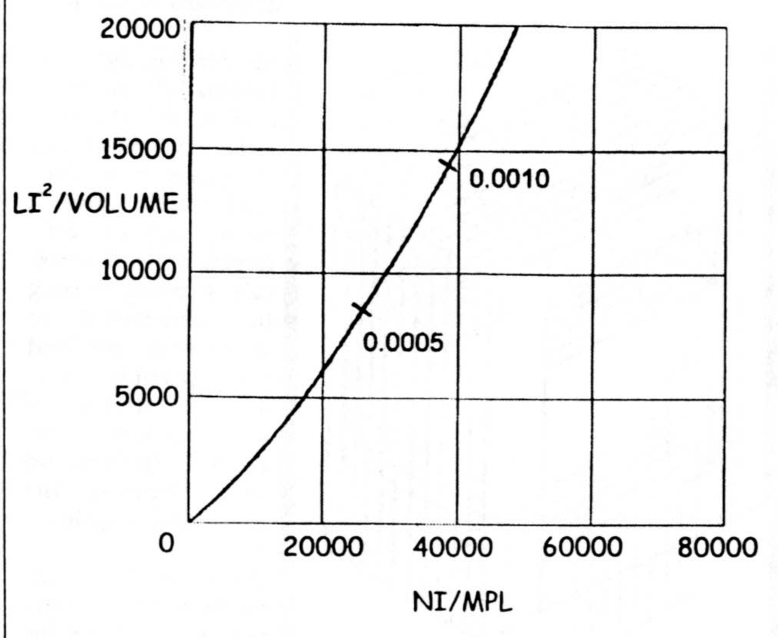

figure 2 - Hanna curve for Silicon Steel, 60Hz, 1 kG

The next step is to determine the number of turns of wire in the primary that will provide 31 henries of inductance when wound around an iron core. Dig through any electromagnetics book and you'll stumble across the inductance equation for an iron core inductor:

$$L = (3.19\ NT^2\ u\ CSA) / (MPL \times 10^8)$$

where:
NT = the number of turns of wire around the core
u = the effective permeability of the core material
CSA = the cross-sectional area of the core in square inches
MPL = the magnetic path length of the core in inches

It seems straight forward enough to rearrange and solve for NT. Armed with a ruler, one can measure the core dimensions to calculate A and MPL.We just need to get a handle on the permeability u. Turns out that u is an elusive animal as you electromagnet heads out there already know. It depends largely on the working conditions and composition of the iron in the OT core. Permeability can be determined from B-H curves, and manufacturers of core products have B-H curves for each type of core material. Unfortunately, I haven't found "scrap iron" as a core material in any of the standard curves. Published ring sample values for permeability range into the tens of thousands or higher, and some transformer designs use guesstimates in the low thousands, but the SE OT has an added degree of difficulty.

The working conditions of an SE OT has a dc bias current that will saturate the core if left unchecked. Saturation means gross non-linearity which is a bad thing for audio. "Gapping" the core cures the dc saturation problem, but greatly lowers the effective u (into the hundreds), which lowers the inductance. More inductance, not less, is desirable for low frequency response. So the consequence of adding an air gap must be compensated for by adding more turns or using a core of larger cross-sectional area or both. Changing the number of turns or the core size changes the working conditions of the iron which changes u which changes.......you get the picture. The cyclical nature of this iterative process is vicious indeed.

At this point it would be really useful to find a relationship involving L and N that is independent of u. Luckily, others have wrestled with this problem before and found a general solution. One such method involves the use of Hanna curves. There are specific Hanna curves for specific core materials. Once again, "scrap iron" does not appear but let's assume that non-oriented 4% silicon steel (Figure 2) is close enough. Use of these curves also requires the volume of the core, the cross-sectional area of the core, and the magnetic path length of the core. See references 1,2,9,11 for Hanna curves.

Figure 3 shows a typical core made up of E-I laminations. One E and one I lamination are paired together to form the flux path for the magnetic circuit. Many E-I pairs are stacked on top of each other to build a core of sufficient cross-sectional area needed to carry the flux. More flux implies a need for more cross-sectional area much like more electrical current implies a larger gauge of wire. It's obvious that a physical break occurs where the flux must travel from the E lamination to its paired I lamination (and vice-versa). This break in the magnetic circuit lowers the effective permeability of the core material. Normally, the E and its paired I lamination are placed as close as possible together to minimize the effect of this break. To further minimize the fringing effects, the orientation of the E-I pairs are alternated as the core is built up. Most power supply transformers and push-pull OT cores are assembled in this way.

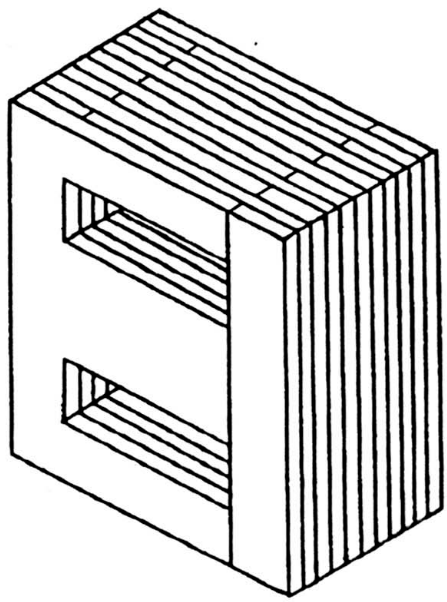
figure 3 - typical core

Iron core inductors carrying a large dc bias require a gapped core. In this construction, the E-I pairs are not alternated. All of the E laminations are stacked on one side and an equal number of I laminations are stacked on the other. The two stacks are placed next to each other but separated by a certain amount of space forming a gap. A spacer of non-ferromagnetic, non-conductive material is used to physically hold the gap to a specific size (Figure 4). So in this construction, the air gap is not actually air at all. Due to the choice of material for the spacer, the gap appears to be air (permeability = 1) as far as the magnetic circuit is concerned.

Figure 5 shows the flux path in an E-I pair. Note that there are two equal, parallel paths. The middle leg of the E lam (tongue) is twice as wide as the outer legs (limbs) to maintain constant flux density in the core. Also note that each flux path crosses the gap twice. The thickness of the spacer is only half of the actual gap length. Power supply chokes (polarized inductors) and SE OTs can be assembled in this way.

All cores using laminations rely on oxide coatings to electrically isolate the laminations from each other. Otherwise eddy currents form, sapping efficiency or in a worse case scenario, causing the transformer to malfunction, overheat and possibly self-destruct. It is vitally important that proper assembly is done to avoid shorting the laminations together.

The width of the middle leg of the E lamination multiplied by the height of the stack of the laminations gives the core cross-sectional area (CSA). Because the core is not a solid material but is made up of a stack of thin sheets, the cross-sectional area is not 100% magnetically useful. A stacking factor (SF) is applied to account for the small, unwanted air spaces that exist between the laminations. Care in assembly, flatness of the laminations, thickness of the laminations, presence of burrs or any foreign objects trapped between the laminations all contribute to the stacking factor. The core cross-sectional area can be calculated using:

CSA = SF x Tongue Width x Stack Height

The scrap power transformer I used yielded a

2 inch high stack of 0.014 inch thick laminations with a tongue width of 1.5 inches. For 0.014 inch thick laminations, I found a reference that used SF = 0.98 , while another one cited an SF = 0.92. Because I'm an amateur stacking used laminations, I set SF = 0.9. The cross-sectional area is:

CSA = 0.9 x 1.5 x 2 = 2.7 square inches

The volume of the core is the magnetic path length (MPL) multiplied by the cross-sectional area of the core:

Volume = MPL x CSA

The MPL is the average length of the path traveled by the flux in one circuit of the core. Assuming the flux path to be rectangular, from Figure 5 it can be seen:

MPL = 2F + 2E + 4D

Assuming the flux path to be a rounded rectangle as drawn in Figure 5, the length becomes:

MPL = 2F + 2E + pi x D

The scrap power transformer core I used has a magnetic path length of 9 inches. Therefore:

Volume = 9 x 2.7 = 24.3 cubic inches

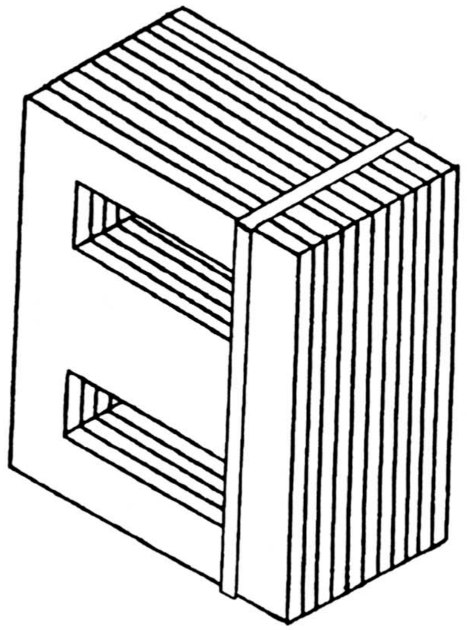

figure 4 - Gapped core

Now to the Hanna curve (Figure 2). The Hanna curve relates the direct current energy stored in the core per unit volume(Y-axis) to the magnetizing field (X-axis). The X-axis is expressed in units of:

NT I / MPL

where:

NT = the number of turns
I = the dc bias in milliamps
MPL = the magnetic path length in inches

The Y-axis is expressed in units of:

$L I^2$ / Volume

where:

L = the inductance in henries
MPL = the magnetic path length in inches
Volume = the core volume in cubic inches

The dc bias current in the OT is the bias current in the output tube. The 845 is biased at 90 milliamps. So

$Y = L I^2/\text{Volume} = 31 \times (90)^2/24.3 = 10{,}300$

Intersecting the curve at Y = 10,300 gives X = 30,000. Rearranging and solving for NT:

NT = X MPL/I = 30000 x 9/90 = 3000 turns

At the curve intersection point the gap-MPL ratio is 0.0007. Therefore the gap length is:

Gap Length = 0.0007 x MPL = 6.3 mils

The spacer thickness is half of the gap length. Spacer thickness is about 3 mils. This size gap maximizes the inductance which would be fine for a smoothing choke. It is important that the OT be linear; the actual gap will be adjusted experimentally.

The transformer turns ratio for 5100 ohms is 25.3 to 1. Therefore, the secondary will have:

Number of turns in secondary = 3000 / 25.3 = 119 turns

It is important to note that these values are for first approximations. The particular Hanna curve in Figure 2 is for silicon steel exercised at low induction which may not accurately represent the SE OT being designed. (The Hanna curve is often used to design smoothing chokes.) When it comes to the actual construction of the coil windings, it may not be possible to get the exact number of calculated turns anyway. As you will see, designing the coil layout will be a juggling of wire diameters, winding layers, and insulation all trying to fit within the core window. It sometimes reminds me of the "measure with a micrometer, mark it with chalk, cut it with a chainsaw" approach. It would be helpful to get a feel for the accuracy of these preliminary design values.

The voltage equation can be used to get a feel for the validity of the transformer design estimates. The flux density B is:

$$= V \times 10^8 \ / \ (25.8 \ FF \ CSA \ NT \ Freq)$$

where:

V = the applied rms voltage
FF = the form factor = 1.11 for a sine wave
CSA = the core cross-sectional area in square inches
NT = the number of turns
Freq = the frequency of the applied signal

Ignoring losses, the power P in the primary equals the power in the secondary (10 watts into 8 ohms). The load reflected into the primary is R = 5100 ohms. So the applied voltage is:

V = sqrt(P R) = sqrt(10 x 5100) = 226 Vrms

The Hanna curve was determined at a frequency of 60 Hz. Solving for the flux density:

$B = 226 \times 10^8/(25.8 \times 1.11 \times 3 \times 3000 \times 60)$
= 1460 gausses = 1.46 kG

The family of Hanna curves for silicon steel have a slope that is directly proportional to the maximum flux density. The particular curve used was plotted for an induction of 1 kG which is slightly less than the calculated value. This means that the approximate design values based on the 1 kG curve shouldn't be too far off, and the error is on the conservative side. An iron core inductor designed and built this way should turn out slightly better (more inductance, fewer turns) than predicted by curve alone.

Thus far it's been determined that 3000 turns of wire are needed for the primary and another 119 turns for the secondary. The next step is to decide how to fit these turns on a bobbin in the space available (window). Also, the physical positioning of these windings (coil layout) will determine the leakage inductance and winding capacitance of the OT. These parasitics are present through out the full frequency range but only begin to show their effects as the frequency rises. Since they can't be eliminated, the goal becomes to smartly arrange the windings in such a way that keeps their effects above the audible range (high frequency rolloff above 20 kHz).

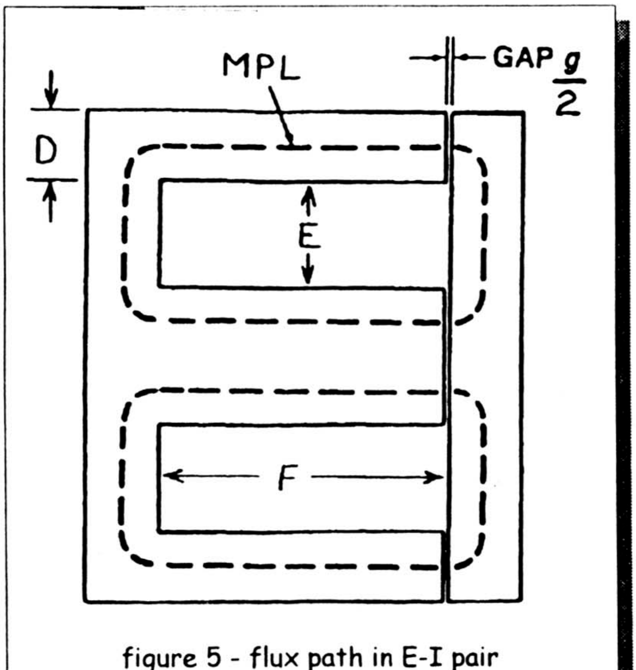

figure 5 - flux path in E-I pair

Next month: taming parasitics, the physical design, and insulation

Reference List

1. "Reference Data for Radio Engineers", Chapter 13 of 6th edition

2. "Practical Transformer Design Handbook", Eric Lowden
3. "Handbook Of Transformer Design And Applications", William Flanagan

4. "Transformers For Electronic Circuits", Nathan Grossner

5. N. Crowhurst, "Electronic Engineering", Leakage Inductance, 4/49, pg. 129

6. N. Crowhurst, "Electronic Engineering", Winding Capacitance, 11/49, pg. 417

7. N. Crowhurst, "Electronic Engineering", Winding Space Determination, 8/51, pg. 302

8. "Radio Receiver Design Vol II", K. R. Stur-

ley

9. "Electrical Engineers' Handbook", Pender & McIlwain, Chapter 3 of the 4th edition

10. Dr. Tom Hodgson, "Sound Practices", To Be, or Not To Be, Linear!, issue 10, pg. 37

11. "Radiotron Designer's Handbook", section on output transformers

12. James Moir, "Glass Audio", Output Transformers, 2/94 pg. 24, 3/94, pg. 30

13. C. R. Hanna, "Trans. A.I.E.E.", Design of Reactances and Transformers Which Carry Direct Current, 2/29

etters

Dan,

I just wanted to yell at you about an article in your October 96 VALVE. In the push pull 6DN7 amp, Dave drops a hint about using an amperite 117N030 relay for wonderful service in delaying the B+ until the fils warm up.
What a shitter... they cost $64.50 each (!) ... ack!
I did some looking and couldn't find any kind of time delay device for under about 45 bucks that could switch 5V@2A.
I'm bitching 'cause I got all excited and it turned out that two relays cost more than the amp itself!
Sigh.
Sorry to rant, but who else would listen?

Tom Ronan, Chicago

Huh?
Well, I guess a guy like Dave, who has a junk pile even bigger than mine, forgets that some of this stuff isn't lying around in everybody's basement.

battery bias and heaters

Aloha Dan:

I just had lithium batteries installed in my C-J PV 12 to set the grid bias, and another battery and charger for the heaters, and then took out the local feedback so it's operating as pure triode.
Talk about moving up to the next level. The noise floor just dropped dramatically. Emerging from the blackness are low level textures and harmonics that are just so natural and relaxed. It's like hooking up megabucks Siltecs, that is, "kinder and gentler", and getting the "high definition" without being aggressive, bright or forward.
That immense amount of fine low level textures and harmonics reminds me of a Joule Electra LA200 (that's where I got the idea of setting the grid bias with a battery). The Joule has a darker, more chocolatey character, with tons of emotional content. Although I haven't heard George Wright's preamps, I bet you lunch that installing batteries will make that unit real scary.

Aloha,
Hiroshi Ito, Honolulu
PS - We also tried that tip in Positive Feedback about putting DC in caps. It works! And it's a killer! Did it on the tweeter crossover circuits for both a Klipschorn and DIY horn system. Wow, the music had a clarity and ease that was just staggering.

I spoke with George about battery bias, that is, replacing the cathode resistors of the 6ER5s with 1.56v batteries (positive to the cathode of the tube and negative to ground). He says go for it in the phono stage of the Wright Phono preamp, but don't do the line stage because the cathode resistor in that stage is run unbypassed to create degenerative feedback, which would be eliminated if the cathode resistor were replaced with a battery.

S.E.X. in Tokyo

Dear Dan,

Your S.E.X. kit amps have served, as promised, as a wonderful introduction to tube electronics. And in fact beat the previously built Kit One amp for low level listening, having better low end. The Kit One, designed around a PC board, didn't really serve as much of a learning experience, although the sonic results proved outstanding.
I spent the first couple of months with the S.E.X. amps running full pelt into a pair of Spendor LS3/5As - not only that, but with only 100 Japanese volts going into the primary for power! They are much more at home with a step-up transformer and Superwhamodynes! Looking forward to bolting on the Magnequest O.T.s. More fun.
Wasn't too struck by the stacked diode mod.
(Continued on page 14)

winding your own single ended output transformer part two

by Jim Flowers

This month Jim discusses parasitics, and the physical layout of the transformer

Parasitics

The parasitic capacitance formed in a multi-layer coil of wire should be easy to visualize. Think of each winding layer as one plate of a capacitor. An adjacent winding layer serves as the other plate, and whatever is between these adjacent winding layers (magnet wire insulation, kraft paper, air space) is the dielectric. The windings will also form a capacitor to ground, to the core, to other windings, electrostatic shielding, whatever. Stray capacitance is a promiscuous fellow; it'll form a circuit with just about anything.

Capacitance is a function of the plate size, the dielectric constant of the insulation, and the distance separating the plates. Therefore, to reduce the interwinding capacitance, the winding layer width should be short, and thick insulation with a low dielectric insulation should be used.

The total self-capacitance of the primary winding is the sum of all the parasitic "capacitors" formed by each pair of adjacent winding layers. These capacitors are in series, not parallel; the total self-capacitance of the primary winding will be less than the value of a single adjacent layer pair.

For example, consider a 3000 turn primary constructed in two different coil layouts. The first layout is two layers of 1500 turns each (a long, narrow coil). The second coil layout consists of 30 layers of 100 turns each stacked upon one another. The second coil will have far less self-capacitance than the first (more than 400 times less). See references 1,2,3,4,6,9,11 for calculations of coil capacitance.

Leakage inductance may be harder to visualize. I think of it this way:
In the ideal transformer, all of the flux produced by the primary winding is intercepted by the secondary winding. In a real transformer, less than 100% of the flux generated actually reaches the secondary. This loss is represented by a parasitic inductance known as the leakage inductance. This leakage inductance occurs because of the physical distance separating the primary winding from the secondary winding. Windings in close proximity exhibit less leakage inductance than windings separated by a greater distance.

The bifilar winding technique uses two wires side by side (they are wound on the bobbin at the same time). Because the windings are separated only by the insulation on their wires, they are very close together, and therefore, there is very low leakage inductance. Unfortunately, this technique has limited use when a high potential difference exists between the wire pair for fear of arc over.

A popular coil layout method, the one that I used, involves inserting layers of the secondary winding in between layers of the primary winding. When layers of one winding (the secondary) is distributed evenly among layers of another winding (the primary), the leakage inductance is reduced because the windings are effectively closer together. This coil winding technique is called interleaving.

Consider two coil layouts for a transformer. In the first layout, start by winding 3000 turns of wire onto the bobbin forming the primary. Put on some required insulation, and then wrap 119 turns of secondary on top of that. In the second layout, wind 1500 turns of wire (half of the primary), insulate, wind 119 turns of secondary, insulate, and finish with 1500 turns on top of that (second half of the primary). The two half primaries will need to be connected together in series to form the total 3000 turn primary.

In the first layout, turn #1 of the primary (at the inner most layer of the coil) is far from turn #119 of the secondary (on the outer most layer of the coil). In the second (interleaved) layout, the secondary winding sits physically between the two half primaries. As compared to the first layout, the whole of the secondary winding is much closer to the whole of the primary winding - it will have less leakage inductance. This interleaving technique can obviously be extended to a greater number of divisions resulting in even less leakage inductance.

In general, reducing the self-capacitance of the winding involves separating the winding layers from each other while efforts to reduce leakage inductance require bringing the windings closer together. Efforts to conquer one problem tends to make the other one worse. As it turns out for the SE OT, leakage inductance is the larger problem, and greater efforts will be applied to its remedy.

Another issue that may not be obvious from the above concerns the total number of turns of wire required for the primary. All else being equal, a transformer that requires 5000 turns of wire will have greater parasitic losses than a transformer that requires only 2500 turns. The 5000 turn transformer will tend to roll off sooner in the high frequencies than the 2500 turn version. Transformers for high plate resistance tubes (211, 811) need higher inductance in the primary to maintain the low frequency response than do transformers for lower plate resistance tubes (2A3, 300B). More inductance implies more turns which implies more parasitic losses

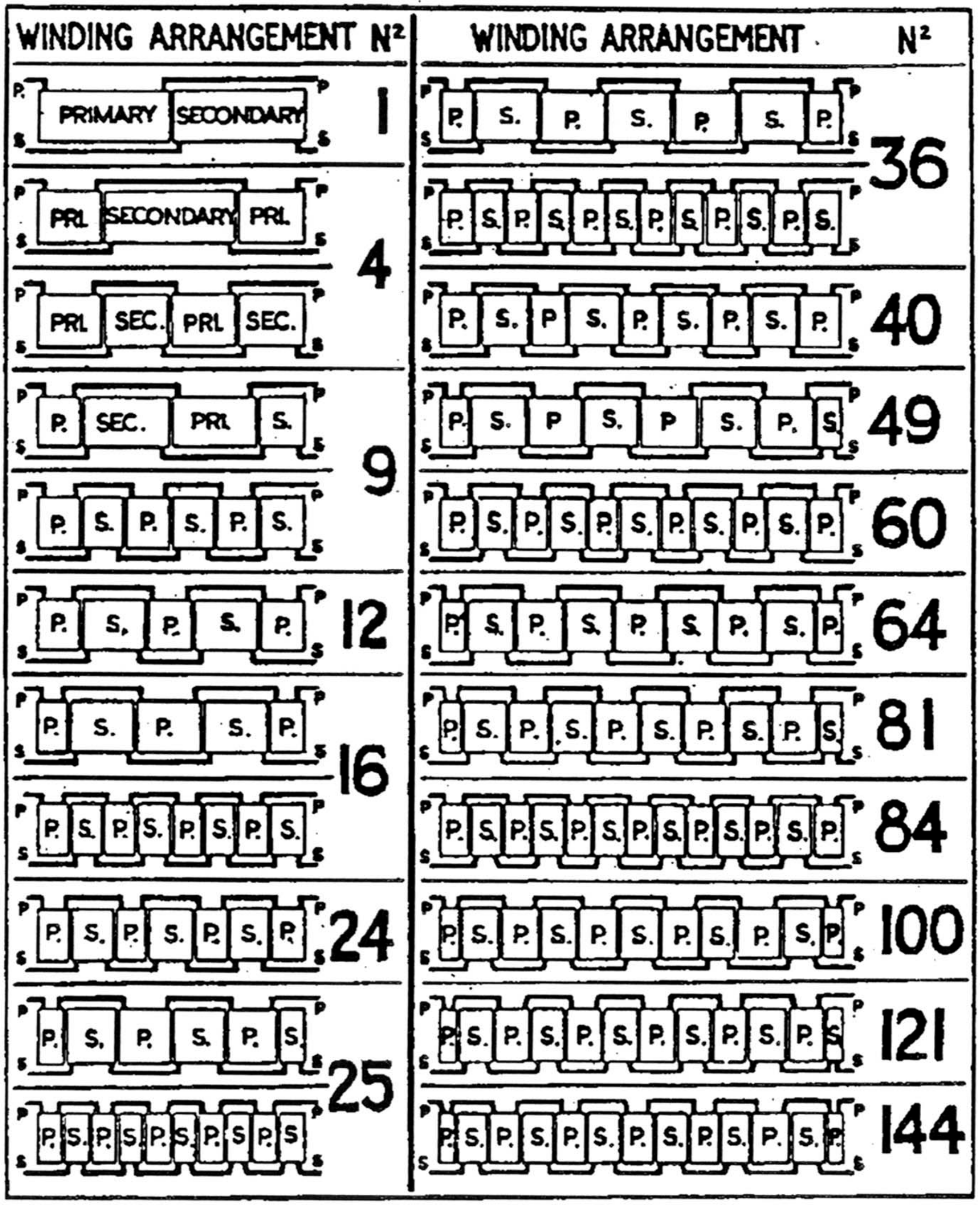

figure 6 - interleave chart

which (you guessed it) results in earlier roll off of the high frequencies. What this means is that high impedance, wideband transformers are considerably more difficult to design and build correctly than low impedance ones. This may also give a hint as to why good interstage transformers are hard to come by.

Here's a thought for you bi-ampers out there. The OT for the bass amplifier doesn't need an extended high end. Wind all the turns you can squeeze in for lots of inductance and extended low end and parasitics be damned! What do you care when it rolls off? The OT for the other amp doesn't need an extended low end. Use fewer turns to better manage the parasitics resulting in an extended high end. Maybe if you get good at this, the OT rolloffs (high and low) can be incorporated as part of the loudspeaker crossover. I haven't tried this yet (or even thought it through completely), but isn't experimenting what this is all about?

Physical Design

Figure 6 shows coil layouts utilizing increasing levels of interleaving. This table of winding arrangements was taken from Crowhurst's excellent article on leakage inductance (see reference 5 in last month's installment). The drawings show cross-sections of only one side of the coil and also have omitted the bobbin. Notice the "N-squared" values assigned to each arrangement. Recognize that the leakage inductance improves (gets smaller) as "N-squared" gets larger. The two coil layouts previously described (leakage inductance examples) are the first two arrangements shown at the upper left of the table. This implies that a 4x improvement resulted from a simple splitting of the primary. The actual improvement also depends on the insulation thickness and a factor of 4 may not be realized. Crowhurst's article provides a chart for calculating the leakage inductance. See references 1,2,3,4,5,9,11 for calculations of leakage inductance.

Examine the two arrangements shown for N-squared equal to 16 (winding arrangements #7 and #8 counting down from the upper left). Layout #8 uses a more complex arrangement (more sections) than layout #7 to achieve the same leakage inductance factor. I think this is due to the improved symmetry of layout #7. The center section of the primary winding in #7 has twice as many turns as its outer sections; whereas, #8 splits the primary into four equal sections. The secondary sections in layout #7 are each sandwiched between two equal sections of the primary; i.e., one-fourth of the primary borders on the outside and another one-fourth of the primary (one half of the center section) snuggles up along the inside. The center section of the primary is equally shared by both secondary sections surrounding it.

The drawings in Figure 6 might appear inappropriate for the SE OT coil layout. After all, the drawings show equal cross-sectional areas for the primary and secondary windings, whereas the SE OT under design has 3000 turns for the primary and only 119 turns for the secondary. But this is actually correct. The secondary while having fewer turns will be wound with a much heavier gauge of wire. For a transformer, power in the primary equals power in the secondary, which means equal cross-sectional areas for equal power densities.

Winding the turns of wire onto the bobbin is done by two methods. In the layer winding method, each turn of wire is neatly placed next to (side by side, not on top of) the previous turn moving across the bobbin in one row. At the completion of the row, a layer of insulation is wrapped around it, and the next layer of wire is wrapped on top of this. (A spool of solder often looks like this minus the insulation of course). The random winding method is less orderly, think of a spool of fishing line. The winding shouldn't been done too haphazardly; a uniform buildup should be maintained throughout the winding process. Insulation is not used between layers (but still used between windings). Subsequent turns will fall into spaces left by the previous turns below.

I used the layer winding method. Although more tedious, I think it is mechanically and electrically more sound. Maybe one day I'll try a random wound version for comparison.

The window is the opening in the core where the turns of wire pass through. Unfortunately, not all of this area is available for copper. Approximately 25% or more of this space is taken up by the bobbin and insulation. (This is different from the true copper utilization factor sometimes referred to as "k" in the textbooks.) Figure 1 (shown last month) shows the dimensions of the available space for the coil build (0.706 inches by 2.125 inches).

Roughly one half of the available window area is used for the primary and the other half is used for the secondary.

Area available for each winding
=75% of (0.7 x 2.125) / 2
=0.558 square inches

For the primary, 3000 turns must fit into 0.558 square inches:

Area per turn of primary
=0.558 square inches/3000 turns
=0.000186 square inches

This available space per turn is shaped like a square. The circular wire diameter that fits in this square equals the length of one side of the square:

Primary wire diameter
= sqrt(Area per turn)
=sqrt(0.000186) =0.0136 inches

According to the wire table in Figure 7 (references 1,2,11) 28 gauge wire could fit. Similarly, for the secondary:

Area per turn of secondary
= 0.558 square inches / 119 turns
=0.00469 square inches

Secondary wire diameter
= sqrt(Area per turn)
= sqrt(0.00469) =0.0685 inches

Again referring to the wire table, there's enough room for 14 gauge wire. This size wire is rather stout and quite challenging to wind. Two or more paralleled, smaller gauge wires could be substituted as an electrical equivalent. A wire gauge rule of thumb: moving 3 numbers in gauge changes the copper cross-sectional area by a factor of 2; i.e., two 17 gauge wires in parallel are roughly equivalent to a single 14 gauge wire. Additionally, four 20 gauge wires in parallel add up to a single 14 gauge wire. These wire splitting strategies allow flexibility in the coil layout design for the secondary winding.

For example, one secondary coil layout consists of four sections each of 14 gauge wire. Each section is a single layer of 30 turns. These four sections are connected in series

Wire Size AWG	Circular Mils	Diameter (in)		Resistance ohms/1000 ft.	Weight lbs/1000 ft.	Random Winding		Layer Winding				Wire Size AWG
		Single Insulation	Heavy Insulation			Single Turns/in'	Heavy Turns/in'	Single Turns/in	Heavy Turns/in	Layer Insulation (in)	Edge Distance (in)	
10	10384		.106	.999	31.7	86	75	9	8	.010	.250	10
11	8226		.094	1.26	25.2	108	95	10	9	.0100	.250	11
12	6529		.084	1.59	20.1	133	130	11	11	.0100	.250	12
13	5184		.075	2.00	15.9	162	159	12	12	.0100	.250	13
14	4109	.0658	.067	2.52	12.6	212	193	14	13	.0100	.188	14
15	3260	.0587	.060	3.18	10.0	255	248	15	15	.0100	.188	15
16	2581	.0524	.054	4.02	7.95	324	316	17	17	.0100	.188	16
17	2052	.0468	.048	5.05	6.32	405	394	19	19	.0070	.188	17
18	1624	.0418	.043	6.39	5.02	525	487	22	21	.0070	.125	18
19	1289	.0373	.039	8.05	3.99	641	596	24	23	.0070	.125	19
20	1024	.0334	.035	10.13	3.16	850	792	27	26	.0050	.125	20
21	812.3	.0298	.031	12.77	2.51	1055	982	30	29	.0050	.125	21
22	640.1	.0266	.028	16.20	1.99	1340	1210	34	32	.0050	.125	22
23	510.8	.0238	.025	20.30	1.59	1370	1260	38	36	.0050	.125	23
24	404.0	.0213	.022	25.67	1.26	1730	1550	42	40	.0020	.125	24
25	320.4	.0190	.020	32.37	1.01	2150	1940	47	45	.0020	.125	25
26	252.8	.0170	.018	41.02	.799	2990	2700	53	50	.0020	.125	26
27	201.6	.0152	.016	51.44	.634	3700	3550	59	55	.0020	.125	27
28	158.8	.0136	.014	65.31	.504	4680	4180	66	62	.0015	.125	28
29	127.7	.0122	.013	81.21	.401	5900	5160	73	68	.0015	.125	29
30	100.0	.0109	.012	103.7	.318	7500	6560	82	77	.0015	.093	30
31	79.21	.0097	.011	130.9	.254	9270	8090	91	85	.0015	.093	31
32	64.00	.0088	.010	162.0	.202	11400	10000	100	94	.0013	.093	32
33	50.41	.0078	.009	205.7	.161	14500	12500	113	105	.0013	.093	33
34	39.69	.0070	.008	261.3	.127	18800	16250	128	119	.0010	.093	34
35	31.36	.0062	.007	330.7	.101	24000	20600	144	133	.0010	.093	35
36	25.00	.0056	.0060	414.8	.0803	29650	25000	158	145	.0010	.093	36
37	20.25	.0050	.0055	512.1	.0641	37400	30900	177	161	.0010	.093	37
38	16.00	.0045	.0049	648.2	.0509	46700	39300	198	181	.0010	.062	38
39	12.25	.0039	.0043	846.6	.0403	62700	51500	226	205	.0007	.062	39
40	9.61	.0035	.0038	1079	.0319	89600	72000	262	226	.0007	.062	40
41	7.84	.0031	.0034	1323	.0252	107800	89800	274	250	.0007	.062	41
42	6.25	.0028	.0030	1659	.0199	133500	116500	304	283	.0005	.062	42
43	4.84	.0025	.0027	2143	.0159	167000	143000	340	315	.0005	.062	43
44	4.00	.0022	.0025	2593	.0127	217000	168500	386	340	.0005	.062	44

figure 7 - wire chart

otaling 120 turns (30 + 30 + 30 + 30).

An alternate layout uses 20 gauge wire. The 20 gauge wire has half the diameter (one-fourth of the area) of 14 gauge wire. Bifilar winding of 20 gauge wire fits 30 turns in one layer just like the single layer of 14 gauge wire but with half the cross-sectional area. To get the same copper area requires another bifilar layer. Combining two bifilar layers in parallel results in four 20 gauge wires of 30 turns each all in parallel. This combination has the same cross-sectional area as the single layer of 14 gauge wire. Four of these double-layer sections are connected in series as before. Note that when connecting sections in parallel, it is important that each section contains the same number of turns.

The first coil layout consists of four layers while the alternate coil layout consists of eight layers. The ability to divide a winding into multiple layers helps make interleaving practical.

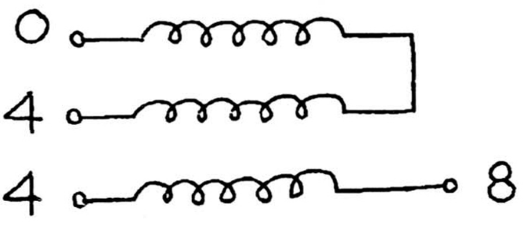

figure 8 - three layer secondary

So far the secondary only has connections for a 8 ohm load. Often, an OT has multiple connections available for different loads. This is accommodated by offering multiple taps on the secondary winding. Depending on the series/parallel combination of taps chosen, the secondary operates as if it has a different number of turns.

Consider the secondary coil layout above that used four sections of series connected 14 gauge wire. If only two sections are series connected (30 + 30 = 60 turns) and this pair is paralleled with the other two sections connected in series (30 + 30), then the total winding becomes a 60 turn secondary ([30 + 30] // [30 + 30]). Since the original 120 turns was for an 8 ohm load, then 60 turns reflects a 2 ohm load.

Figure 8 shows a 3-layer secondary section consisting of three layers of 40 turns each. When all three layers of the section are connected in series (jumper the 4 ohm terminals together, speaker load on terminals #0 and #8), the 120 turns reflect an 8 ohm load. Connecting the speaker load to terminals #0 and #4 exercises only 80 turns of the secondary reflecting a 4 ohm load (3.6 ohms actually). When connecting sections in series, it is important to observe correct polarity to avoid bucking the windings.

With this secondary coupled to a 3000 turn primary, the transformer could be identified as a 5 kohm OT with a 0 - 3.6 - 8 ohm secondary, or a 5.6 kohm OT with a 0 - 4 - 9 ohm secondary.

Another consideration is the ohmic resistance of the wire that make up the windings. Less resistance is better, but that means thicker wire. Thicker wire might not fit in the allowed space without reducing the total number of turns. Increasing the allowed space requires different laminations having a larger window area.

A goal for the ohmic resistance of the secondary is one percent of the load it drives. Too much resistance adversely affects the insertion loss and damping ratio. For an 8 ohm secondary, the goal is 80 milliohms or less. The wire table provides resistance per foot for each gauge of wire. To calculate the length of wire in the winding, multiply the number of turns by the average length per turn (ALT). Determine the average length per turn in the same way that the MPL was calculated.

ALT = 2 x Tongue Width + 2 x Stack Height + 4 x Coil Build

Substituting:

ALT = 2 x 1.5 + 2 x 2 + 4 x 0.7 = 9.8 inches

Because the shape of the windings is more of a rounded rectangle, an alternate formula is sometimes used:

ALT = 2 x Tongue Width + 2 x Stack Height + pi x Coil Build

Substituting:

ALT = 2 x 1.5 + 2 x 2 + 3.14 x 0.7 = 9.2 inches

And as if that wasn't enough, according to a laminations datatable (see reference 3), the average length per turn of a 2 inch high stack of EI-13 laminations is 9.63 inches.

The resistance of 120 turns of 14 gauge wire at 9.63 inches perturn and 2.52 ohms per 1000 feet:

Rsec = 2.52 x 120 x 9.63 / 12000
= 0.243 ohms

This value is three times greater than the stated goal. The wire's resistance is a function of length (turns and length per turn) and width (gauge). To get to 80 milliohms, the secondary can have three times fewer turns, or three times as many sections to add in parallel. Using only 120/3 = 40 turns in the secondary requires the primary to be cut to 3000/3 = 1000 turns to maintain the same transformer ratio. The primary inductance is now lowered drastically. To make up for the lowered turns count, the core cross-sectional area can be increased by adding more laminations to the stack. The larger cross-sectional area results in a greater length per turn which partially offsets the decrease in winding resistance achieved by using fewer turns.

To get a three fold increase in wire gauge will require three times the allotted window area reserved for the secondary wiring. Whether this is accomplished by using thicker wire or more sections in parallel, the window must be larger, which means a larger set of laminations. It is not good practice to increase the area used by the secondary at the expense of the primary.
The ohmic resistance of the primary winding contributes to the insertion loss as well. If the insertion loss is to be split equally between the primary and the secondary, then the primary winding resistance should equal the reflected secondary winding resistance. The Radiotron Designer's Handbook suggests a primary winding resistance of twice that value. (I don't understand why it should be twice the value.) Based on the ambitious target of 80 milliohms of secondary winding resistance, the primary winding resistance should be:

Rpri = 2 x N^2 x Rsec = 2 x $(25.3)^2$ x 0.080 = 102 ohms

If both targets are met, then the total loss due to winding resistances is only 3%. When 97% of the power passes through the OT, the insertion loss is only 0.13 dB.

Realistically, the maximum copper insertion loss should be well below 0.5 dB. While half a decibel doesn't seem like much, this is equivalent to a 10% power loss that only serves to warm up the OT. In other words, only 9 watts actually reach the loudspeaker from an output tube pumping out 10 watts.

Also, the copper losses were ignored when the turns ratio was determined. This results in an impedance error equal in magnitude to the power loss. The turns ratio will appear higher than when measured under the no-load condition.

For a point of comparison, resistance readings taken from several commercial SE OT's reveal an approximate range of 120 ohms to 400 ohms in the primary, and 0.1 ohms to 0.8 ohms in the secondary. Remember that these are guidelines and design goals.
If an OT that you are presently using measures outside these ranges and still sounds good, then don't sweat it. Don't presume the worst without first taking a listen.

Either way, lowering the insertion loss results in a larger transformer. A larger transformer coil will have greater parasitics causing an earlier rolloff in the high frequencies. Compensating for the rise in parasitics requires a more complex coil which could mean more interleaving which adds to the total insulation thickness in the coil which requires a larger window, ad nauseum.

Keeping losses low and fidelity high in the SE OT requires a large, carefully designed, and painstakingly assembled transformer. This becomes even more critical when both the transformer impedance and dc bias current go higher. Good engineering practice comes at a price. This is another reason why the best transformers cost several hundred dollars or more.

Well, we're about halfway there, bottleheads. Next month we'll pick up with insulation, choosing a winding layout, and discussion of building the coil.

Doc B.

winding your own single ended output transformer part three

By Jim Flowers

Insulation

The wire table in Figure 7 (*see Feb '97*) suggests layer insulation thickness. This is the insulation used when winding in the layer method (not random wiring). This is the insulation between layers of an individual winding (layers of the primary), not between windings (where primary meets secondary). The thickness of insulation separating adjacent windings depends on the voltage potential between those windings - the higher the voltage, the thicker the insulation. The minimum thickness also depends on the dielectric strength of the insulation. Materials that can withstand higher voltage gradients don't need to be as thick.

Kraft paper is often used as insulation. A good quality paper can also be used. The paper needs to be uniform and free of acid. Over time the acid in the paper can corrode the wire's insulation. Electrically, multiple layers of thin paper are better than a single thick layer. I have used wax paper (right out of the kitchen drawer) as insulation.

I'm not a materials expert, but I'm sure there are better, more modern materials available. The insulating material should have a low dielectric constant to minimize the parasitic capacitance. The material must maintain its thickness over time and not cold flow. Preferably, the material is non-hygroscopic, i.e., it won't absorb moisture. Paper fails this test, but others such as mylar, polypropylene, and kapton, do not. "Transformers For Electronic Circuits" (reference 4) has a particularly good chapter on insulation. This is an area still under investigation for me, so anyone knowledgeable on the subject please speak up.

Choose A Winding Arrangement

In order to decide which winding arrangement from Figure 6 to use (*see Feb '97*), the maximum tolerable amount of leakage inductance and parasitic capacitance has to be determined. Figure 9 shows a simplified equivalent circuit for a step-down transformer at high frequencies. Rp is the plate resistance of the output tube, C1 is mainly the self-capacitance of the primary, Llk is the leakage inductance, N is the turns ratio, and R is the load resistance.

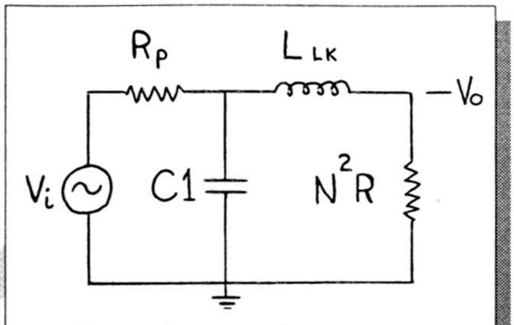

figure 9 - equivalent circuit

Ignoring C1, the transformer equivalent circuit simplifies to a series low-pass LR circuit. R is the series combination of the output tube plate resistance and the load resistance reflected into the primary. The -1db point occurs where the reactance of the leakage inductance is equal to one half of the series resistance.

L =
0.5R/[2 x pi x (Frequency at -1dB)] = (1700 + 5100)/(4 x pi x 20,000) =
27 mH

This time ignoring Lp, the transformer equivalent circuit simplifies to a series low-pass RC circuit. R is the parallel combination of the output tube plate resistance and the load resistance reflected into the primary. The -1db point occurs where the reactance of the parasitic capacitance is equal to twice the series resistance.

C = 1/[2 x R x 2 x pi x (Frequency at -1dB)]
= 1/[4 x (1700 // 5100) x pi x 20,000]
= 3.1 nF

As mentioned before, the interlayer capacitance is a function of the layer's surface area and separation, and the dielectric constant of the insulation. For a single pair of wire layers, the parasitic capacitance Co is:

Co = 0.225 DC AMT Wb / d (in pF)

where:

DC = the dielectric constant of the insulation

AMT = the length of winding turn in inches

Wb = the width of the bobbin in inches

d = the thickness of the insulation in inches

Using a dielectric constant of 3, an average turn length of 9.63 inches, a bobbin width of 2.125 inches, and an interlayer insulation thickness of 1.5 mils:

$Co = 0.225 \times 3 \times 9.63 \times 2.125 / .0015$
$= 9.2$ nF

The primary winding is built up from many layers. The self-capacitance of the primary winding Cp is calculated as follows:

$$Cp = 4/3\ Co\ (NLp - 1) / NLp^2$$

where:

NLp = the number of layers in the primary

Assuming the primary winding is made up of 20 layers, the primary self-capacitance is:

$Cp = (4/3) \times 9.2\text{nF} \times (20 - 1) / (20)^2$
$= 580$ pF

In the simple, second-order transformer model above, C1 equals Cp. Because C1 is much less than the tolerable maximum (580 pF < 3.1 nF), the model predicts that the self-capacitance of the primary will not cause a frequency response rolloff in the audio band.

Leakage inductance is also based on coil geometry. It is approximated by the following formula:

$$Llk = k\ IF\ NT^2\ ALT\ [d + dCu/3] / Wb$$

where:

$k = 3.19 \times 10^{-8}$ (when all dimensions are in inches)

IF = the interleave factor = 1 / "N-squared"

NT = the number of turns

ALT = the average length of a turn

d = the sum of all insulation spacing between sections

dCu = the total copper thickness

The simplest winding arrangement of one primary section covered by one secondary section has no interleaving and therefore "N-squared" = 1. (See upper left corner of Figure 6.) The transformer model is referred to the primary, so NT = 3000. There is only one interwinding insulation space. The interwinding insulation thickness is 10 mils, so d = .01 inches.
Estimate the copper thickness at 75% of the window height making
$dCu = 0.75 \times 0.7 = 0.5$ inches.

$Llk = 3.19 \times 10^{-8} \times 1 \times (3000)^2 \times 9.63 \times [0.01 + 0.5/3] / 2.125$
$= 229$ mH

This is more than the allowed 27 mH. Interleaving will be needed to lower the leakage inductance. The leakage inductance is too high by a factor of 8.5. Try winding arrangement #7 having an "N-squared" value of 16. (Arrangement #6 with "N-squared" equal to 12 would probably be sufficient, but #7 has a better interleave factor with the same coil complexity.)

$Llk = 3.19 \times 10^{-8} \times (1/16) \times (3000)^2 \times 9.63 \times [(4 \times 0.01) + 0.5/3] / 2.125$
$= 17$ mH

Using this winding arrangement, the calculated leakage inductance is less than the maximum tolerable limit determined above. It is best to have some headroom because leakage inductance doesn't act alone as assumed in the first-order analysis of the model. When doing the coil layout design, the distances used for insulation and copper thickness will be determined more accurately. Smarter analysis and implementation of the transformer model will give better results.

Coil Layout Design

Designing the coil layout is a trial and error procedure. Choose a winding arrangement, select the wire gauges, the insulation thickness, add it all up, allow for some margin, and check to see if it all fits in the allowed space. Compromises have to be made. Yours will be different from mine. Here's what I did.

I started with the secondary layout. I felt that 120 turns resulted in too much resistance. The available space wouldn't allow thicker wire, so I chose to use fewer turns in the secondary. Figure 10 shows one of the four sections. As shown in the figure, each section is made up of 2 layers. The first layer is 54

(Continued on page 8)

turns of 19 gauge wire. The second layer consists of 27 bifilar turns of 19 gauge wire. The layout shown has six terminals to allow

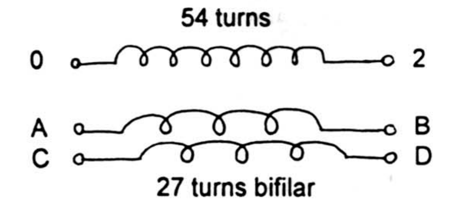

fig. 10a - 2-layer secondary section

flexibility to experiment with many different load connections and combinations. It is interesting to see the differences in the frequency response of the same partial winding in each of the four sections. For example, the 54 turn layer in section #1 rolls off in the high frequencies differently from the 54 turn layer in section #4. The astute coil parasitics experts in the audience will knowingly attribute the differences to the physical location of the winding within the coil.

Dispense with the flexibility and consider the entire secondary winding as shown in Figure 10a. In each of the four sections, the terminal pairs 2-to-A and B-to-C have been permanently wired together.

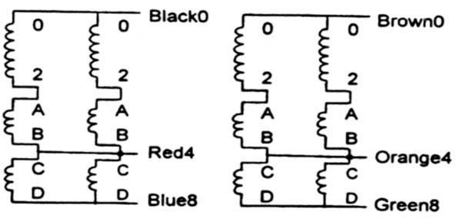

fig. 10b - complete secondary

The 8 ohm connection is a 108 turn secondary (54 + 27 + 27). This requires two jumpers; one connecting Black0 to Brown0, and another connecting Blue8 to Green8. The 8 ohm load is connected across Black0 and Blue8.

The 4 ohm connection (actually 4.5 ohms) is 81 turns (54 + 27). This requires two jumpers; one connecting Black0 to Brown0, and a second one connecting Red4 to Orange4. The 4 ohm load is connected across Black0 and Red4.

The 16 ohm connection is 162 turns [(54 +

27) + (54 + 27)]. This requires one jumper; connect Red4 to Brown0. The 16 ohm load is connected across Black0 and Orange4.

Changing the number of turns in the secondary forces a change in the number of turns in the primary to maintain the transformer ratio. To keep the "transformer impedance" at 5.1 kohm, the primary must have:

NTp = N NT = 25.3 x 108 = 2732 turns

This is somewhat less than the 3000 turns originally specified. The inductance equation discussed earlier shows inductance to be a function of the turns ratio squared, so dropping to 2732 turns from 3000 would seem to result in a large drop in inductance. In practice, the permeability u also changes; and the inductance in an SE OT behaves more like a linear relationship with respect to the number of turns rather than a square law. The Hanna curve can be used to estimate the new inductance:

X = NT I / MPL = 2732 x 90 / 9 = 27320

According to the curve, for X = 27,000 the Y value is 9,300.
Solving for L:

$L = Y\ Volume / I^2 = 9300 \times 24.3 / (90)^2 = 27.9$ H

Going back to the wire table, 140 turns of 28 gauge wire will fit in one layer across a 2.125 inch wide bobbin. I actually fit 144 turns per layer, perhaps the wire table is conservative. Using 144 turns per layer, 19 layers are needed to reach 2700 turns.

Figure 11 shows the coil layout. Note that this winding arrangement does not strictly follow the interleave pattern. The outer-most primary sections are not precisely equal to one half of the inner sections. To follow the pattern exactly requires the outer-most primary sections to have 2.5 layers instead of only 2. There wasn't enough room for the half-layers; they were omitted and consequently, an unknown (probably small) increase in leakage inductance resulted.

The wiring shown is only representative; it is not the exact connections. The intent is to show that the primary sections are connected in series, and the secondary sections are connected in parallel. The secondary requires

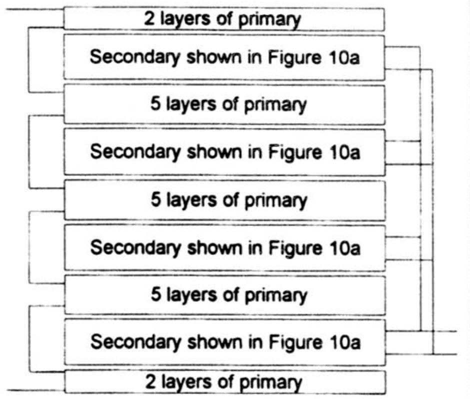

figure 11 - coil layout

additional jumpers that depend on what load will be attached.

It might seem reasonable (or is it heresy) to provide a center tap on the primary so that the transformer can be used with a push-pull connection. The center tap lead would be attached at the 72 turn of the third layer of the central "5 layers of primary" splitting the primary into two equal parts. Perhaps you recognize that the parasitics of the outer half of the primary are not equal to those of the inner half due to the physical coil geometry. This results in an imbalance that only gets worse as the output stage begins to operate in class B. Be aware that the best coil layout for push-pull is different than shown here. All of the textbook references have examples for distributing the primary winding for push-pull operation.

Coil Build

The wax paper interlayer insulation measures 1.25 mils thick. Insulation between windings is 10 mils thick. There are 19 layers of 28 gauge wire plus 8 layers of 19 gauge wire plus 19 pieces of interlayer insulation plus 8 pieces of interwinding insulation.

Coil Build = 19 x 0.0136 + 8 x 0.0373 + 19 x 0.00125 + 8 x 0.010 = 0.661 inches

The height of the available window measures 0.706 inches. Therefore, the left over space is:

0.706 - 0.661 = 0.045 inches

(Continued on page 10)

(Continued from page 9)

This is not much room left over. Some space must be allowed for the inevitable bulge in the wire. According to Lowden (reference 2), allowing for 15% is common practice. That much extra puts the coil build over the limit. Previous experience with this size coil winding led me to try anyway, and I was able to make it fit, but just barely. A lot of care was taken to ensure that the windings were wrapped snugly and that no slop built up. I was tempted to reduce the interwinding insulation thickness below 10 mils to make some extra room. As I am not confident about the insulation requirements at this time, I retained the full 10 mils thickness.

At this point, the easy part is done. Theory, design, and the math to back it all up is easy once you get used to it. Winding on the other hand...well...I'm still not used to it.

The Winding Machine

It is necessary to fashion some sort of a winding machine (see Figure 12). I used half-inch threaded rod as a spindle, plywood for a base and two pillow blocks, poplar wood for a mandrel and the crank handle, and lots of nuts and washers to hold it all together. I also constructed a separate wooden base to hold the wire spool in place (horizontally) during the winding.

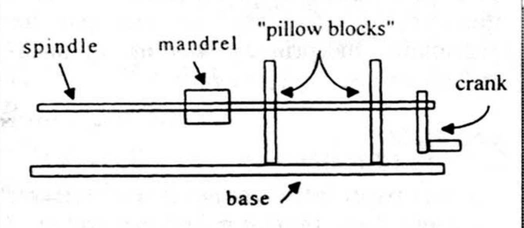

figure 12 - coil winder

I didn't have a turns counter. Only the secondary turns count is critical (during parallel connections) and the 54 turns of 19 gauge wire was easy to count. I counted a single layer of 28 gauge wire just once to verify the turns count per layer in the primary.

It is important for the mandrel to securely hold the bobbin. The mandrel was formed from two pieces of 0.75 inch thick poplar cut to size on a table saw and then screwed together (see Figure 13). It required a half-inch hole along its central axis to fit on the threaded rod spindle. It was easy to center a square pilot hole using the rip fence on the table saw. A half-inch drill bit enlarged the hole to proper size and shape.

The spindle extends out far enough to carry along the 28 gauge wire spool while winding the 19 gauge secondary. The spool is tightly clamped to the spindle so that it rotates with the bobbin assembly. I did this so that I wouldn't have to splice the primary wire across each secondary section. This really isn't worth the hassle and not particularly recommended.

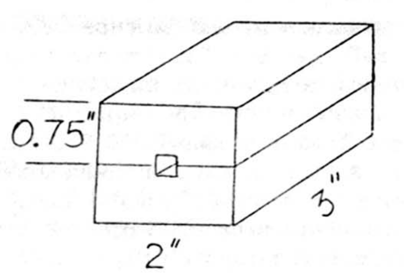

figure 13 - mandrel

As you can tell, this winder is very crude. It did work adequately though. I'm sure you can do much better.

Winding

The bobbin I used is a rectangular tube with flanges, also called end plates or cheeks. The flanges provide the margin (insulating distance) at the sides of the coil and would support the coil if random wiring was used. If the bobbin didn't already have flanges, I would add them because I'm not certain that my winding technique is good enough yet. I'm concerned that the first and/or last turn of an upper layer might slip off the edge and fall down to a lower layer. Even with the help of the flanges, I had to be careful to avoid this mishap.

The squarish bobbin has two exposed faces and two interior faces. The interior faces fit into the core window between the tongue and the limbs of the E lamination. Take extra care to keep the winding layers flat and as compact as possible on the interior faces. Confine all coil maintenance to the exposed faces. This includes leads, splices, crossovers, taps, tape anchors to secure the start and finish points of each layer, and insulation overlap.

For thin wire, like 28 gauge, use a flexible wire as a lead at the start and finish of the winding. Splice and tape anchor the lead at an exposed face. Figure 14 shows one method of anchoring the first turn of wire.

Lay each turn down neatly side by side until the other side of the bobbin is reached. Tape anchor the last turn. Wrap a layer of insulation and begin the next layer. The style of layer winding is defined in two ways depending on the polarity of adjacent windings. In a conventional winding traverse, also called a U winding, the polarity switches at each layer. In a flyback winding traverse, also known as a Z winding, the polarity of each layer is the same because the rotation and direction of the winding traverse are the same from layer to layer. The Z winding requires that the first turn of each layer starts on the same side (at the same flange) of the bobbin. See Figure 15 for a pictorial representation of the U and Z winding methods.

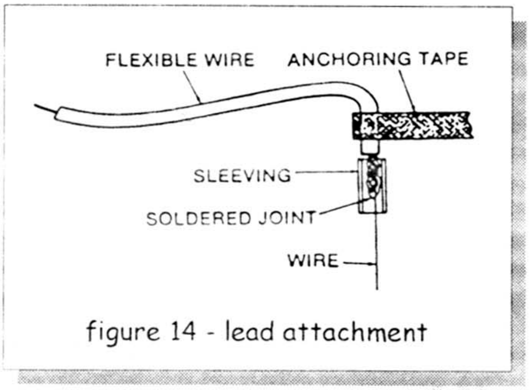

figure 14 - lead attachment

Using a flyback winding traverse minimizes the peak voltage difference between adjacent layers in a coil. This reduces the stress on the insulation and also lowers the effect of the interlayer parasitic capacitance. Less capacitance extends the upper bandwidth pushing the rolloff to a higher frequency. Refer to Grossner (reference 4) for a detailed explanation.

For me, the entire coil winding process was spread out over a period of five evenings. Keep a written record of the build progress. It's not encouraging to resume the next day uncertain if this newest layer is number 4 or number 5.

Cutting the paper insulation to the exact width of the bobbin can be tricky. Stiff paper can be accurately cut on a paper cutting table - those curved one armed slicing things. Thinner materials require a straight edge and a razor blade. The best cutting surface I've found for the thinnest insulation is the Rotary Mat made by Olfa that my wife uses for cutting patterns in clothing material. The 2 ft by 3 ft mat has a grid on it marked in inches and the cutting tool resembles a rolling-wheel pizza cutter. Rolling the cutter along a steel ruler gives straight, accurate cuts.

After all winding is finished, wrap the coil in a protective layer of insulation. If solder tag terminals are to be attached to the coil, heavily spot insulate the coil at this location. Solder the winding section leads to the terminals per the layout paying careful attention to polarity. If you are not completely sure which wire is which, this can be determined with a signal generator and an ac voltmeter later. The secondary I used required 24 connections to 6 different terminals. Hopefully, the insulation coating the magnet wire is the soft solderable type - it will melt away during the soldering process exposing the copper beneath (*this type is polyurethane insulation, not the "high temp" stuff - Doc B.*). Other types of coatings must be scraped off. Of the soft solderable types, I was told that the red colored coating is easiest to work with, followed by gold, and then green, and lastly blue.

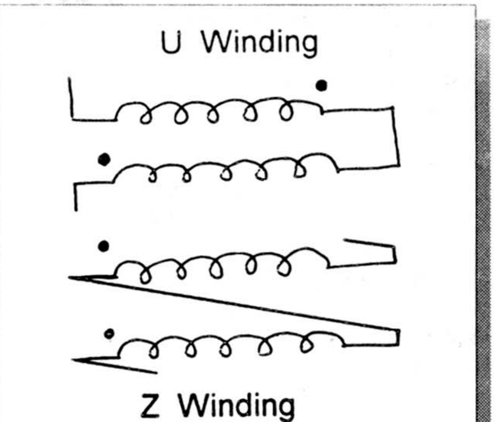

figure 15 - winding styles

Transformer Assembly

Now its time to gather up the laminations and form the core. Before starting, be sure that the screw holes punched in the laminations are clean. If the salvaged core was impregnated, the residual potting compound on the laminations can be used to reduce transformer "singing" due to magnetostriction. Stack the E laminations together forming the E-core. Sandwich the E-core between two flat steel plates and clamp the assembly together with C-clamps. Carefully square all sides of the core and keep the edges of the laminations flush particularly where the gap will be formed. When all is square, tighten the clamps. I used my table saw as a jig to help square the core and clamp assembly. Follow the same procedure for the I laminations.

Let the core soak in a hot oven until the potting compound has softened throughout. Use the temperature and time learned from disassembling the core (probably around 300 degrees for about 30 minutes). Remove the core pieces from the oven and set on a flat surface to cool. Ensure that everything is still square; the softened compound may allow the stack to shift.

Remove the clamps after the core has cooled down. The steel plates come free from the core because the lamination surfaces in contact with the plates had been cleaned with solvent before clamping.

Fit the bobbin onto the E-core; this should be a tight fit. I don't recommend using the keepers (assuming they survived core disassembly). If the keepers are absolutely required to fit the core snugly into the bobbin, they must first be modified. Trim the legs so that they are slightly shorter than the legs on the E laminations. It is vitally important that no ferromagnetic material crosses the air gap.

Add the spacer and the I-core and bolt it all together with the appropriate hardware. Make sure all screws, brackets, and even the end bells are of non-ferromagnetic construction; use brass for example. It would be irresponsible to build such a nice transformer with hardware that magnetically saturates with every beat of the music.

OK bottlebuds, we're done building. Next month Jim wraps it up with testing and adjusting our creation - Doc B.

letters

see, we don't just talk about SE

Hi Dan,

Got the three years back issues. Will try to catch up with you guys.
Built the 4 watt PP 6DN7 according to VALVE October '96, with full wave bridge rectifier and cap regulator. With two 330 mfd caps and a 0.1 mfd coupling cap, the sound is too bassy (no feedback) with my dad's old ITT speakers (+25 yrs.). Changing the B+ caps to two 150 mfd sounds great.
The 120 mA stereo rating for B+ is underrated. With a measured 305VDC, the bias for each 6DN7 is found to be about 40 mA, so the custom order power transformer (250V, 150mA, 6.3V, 5A) is running hot.
Yes, here in City of Angel, you can order (custom made) one 70W power transformer for $16 and two 10K ohm CT 10W output transformers for $10 each. Since the 6DN7 is now running at about 12 watts above the rated 10 watts, what would be the best solution, reduce the B+ or increase the cathode resistor value?
Waiting for EF86 and 6A3 delivery from AES. The custom order output transformer for this project - 25W, airgapped iron with 3-4 mm separation, with paper or plastic - price? $20 each.

T. Song
Bangkok, Thailand

PS - If you don't let the blue smoke out some time, you will get blue all over...

T. - My inclination would be to increase the cathode resistor value on the PP 6DN7 so you run your power transformer cooler, along with your tubes. Doc B.

S.E.X.tasy with Janis Joplin

Doc B.,

Hi, I finally managed to finish building the S.E.X. amp this past weekend. What took me longest was finding the 14 ga, solid copper wire for the ground buss. The nearest place that sold it, when I allowed my fingers to do the walking for me, was a good 50 minute drive away.
Then one day I happened upon a moving sale from a local hardware store and there I got a

winding your own single ended output transformer -conclusion-

By Jim Flowers

Test and Measurement

The size of the air gap greatly affects the operating conditions in the SE OT because it dominates the reluctance of the magnetic circuit. It's been argued that this degree of dominance renders the type of core laminations to be of little importance - nickel, permalloy, vanadium, or even kryptonite, it makes no difference. Although this conclusion is based on the magnetic formulas, I guess it is up to the listener to judge. After all, the electrical formulas certainly don't explain why some listeners prefer silver wire over copper wire. Either way, I'm sure a lot of great transformers have been made using the
less exotic alloys and copper wire.

The presence of the air gap "linearizes" the transformer iron. Read Hodgson's article (reference 10) for a detailed explanation of the gap's effects on transformer operation. A larger gap allows a larger dc bias and a more intrinsically linear transformer, but this comes at the expense of lower primary inductance and increased magnetizing current. Allowing the magnetizing current to grow excessively changes the straight load line into a rounded load ellipse. The size of the gap-needs to be tweaked for best performance.

The graph in Figure 16 shows how primary inductance changes as a function of gap length. There is a gap size that maximizes the inductance for a given standing current. All else being equal, a larger bias current requires a larger gap and also results in less inductance. This is the gap size suggested by the Hanna curve.

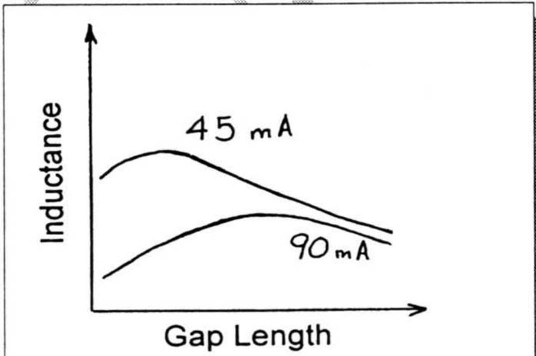

Fig.16 Inductance vs. Gap Length

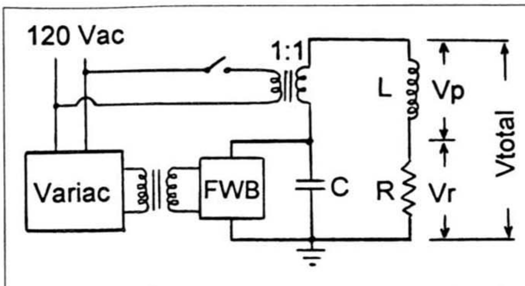

Fig. 17 Inductance measurement circuit

Figure 17 shows a test setup to measure the primary inductance. A variac controls the input voltage to a full-wave bridge circuit that supplies a dc voltage to the transformer under test. This dc voltage setups up the standing current (dc bias) in the primary. The standing current is measured by the dc voltage drop across resistor R. Capacitor C has a large value to minimize ripple. A 1:1 transformer applies a 120 Vrms 60 Hz excitation signal to the circuit. The resistance of the primary winding Rpri (not shown) is part of the inductor L. Measure Rpri with an ohmmeter before connecting the transformer in the circuit. A voltmeter measures the ac signals Vp, Vr, and Vtotal as drawn in the diagram.

Dial up the desired standing current using the variac and switch in the 60 Hz excitation signal. Measure the ac voltages Vtotal, Vp, and Vr. The voltage Vtotal should remain constant except for line sags in the main power. Therefore, to do the inductance calculation, only Vr needs to be measured at each change in the bias current. Note that according to the voltmeter, Vtotal does not equal Vp + Vr due to the phase differences. It does add up correctly when using vector addition accounting for phase.

To calculate the primary inductance:

$$Zt = R\ Vtotal / Vr$$

$$XL = w\ L = \mathrm{sqrt}[\ Zt^2 - (Rpri + R)^2\ ]$$

$$L = XL / w = XL / (2 \times pi \times 60)$$

For best accuracy, R should be approximately equal to the primary reactance XL. Going back to the design calculation for the primary inductance, the target XL is approximately:

2 x (1700 // 5100) = 2550 ohms

Using R = 2550 ohms at 90 mA drops 230 volts at 21 watts. Using half the suggested R value lowers the wattage requirement to about 10 watts with only a small penalty in accuracy.

Making use of the amplifier, instead of the setup in Figure 17, I inserted a 1250 ohm 10 watt resistor between the 845 plate and the SE OT primary. The grid voltage was adjusted to compensate for the large drop in plate voltage (112 volts) such that the bias current was kept at 90 mA. When the amplifier was driven with a signal generator I was able to measure the inductance at frequencies other than 60 Hz. The measured inductance changed little with frequency. This consistency verifies that using a 60 Hz drive signal in the test setup of Figure 17 gives a reasonably accurate measure of the primary inductance.

For those of you with a beefy ac power supply and/or a high voltage dc source, feel free to substitute the appropriate box for the discrete parts used in the test setup.

If the test is run several times while incrementing the gap size each time, the relationship predicted by the graph in Figure 16 can be seen. The maximum inductance occurs when Vr is at a minimum voltage or Vp reaches a maximum.

When changing the gap size, turn off both the ac and the dc sources. I oriented the transformer on the bench so that the I-core is on top

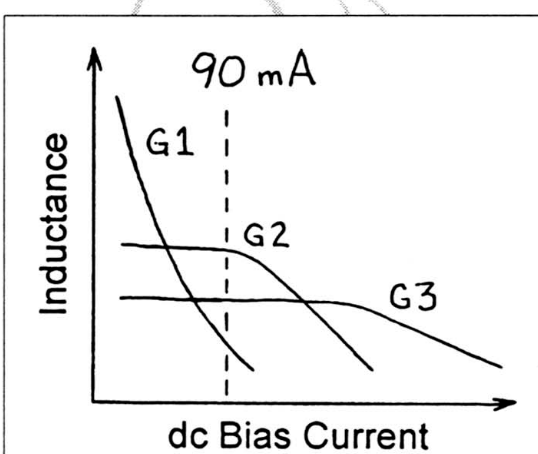

Fig. 18 Inductance vs. Bias

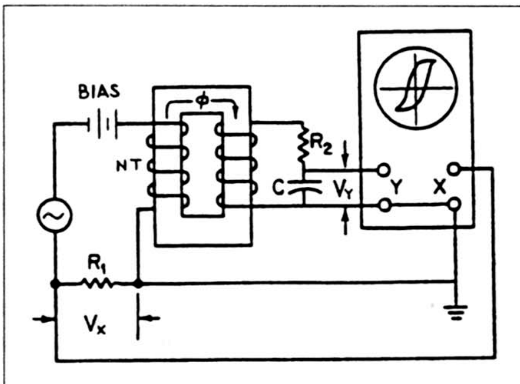

Fig. 19 B-H curve display circuit

to make gap changes without disconnecting the OT from the circuit. Before tightening the screws, apply a small amount of dc voltage. This will magnetically pull the I-core to the E-core, holding the transformer together while you tighten the screws.

Using the test setup in Figure 17 with R = 1242 ohms, Rpri = 136 ohms, and Vtotal = 123.8 Vrms, Vr reached a minimum voltage of 11.48 Vrms at a gap size of 7.2 mils. The maximum inductance is:

Zt = 1242 x 123.8 / 11.48 = 13.4 kohm

XL = sqrt[$(13.4k)^2$ - $(136 + 1242)^2$]
= 13.3 kohm

L = XL / w = 13.3k / (2 x pi x 60) = 35.3 H

This is a little larger than the targeted 31 henries as expected.

A well designed SE OT will exhibit the trend indicated by the plot in Figure 18. Below a certain threshold of bias current (depending on the gap size), the measured inductance remains reasonably independent (constant) of the change in bias current. Above the threshold, the inductance tapers off. If the gap is too small, the inductance will rise dramatically at lower bias currents. When the gap is large, the inductance remains constant over a larger range of bias currents, but at a lower overall primary inductance.

This can be seen with the test setup of Figure 17. Keep the gap size constant, and monitor the voltage Vr while sweeping the bias current (variac). Try this at different gap sizes and the trend will emerge.

The test setup in Figure 19 can be used to display the B-H curve on a oscilloscope. To see the classic B-H curve hysteresis loop, connect a regular power supply transformer in place of the SE OT (remember to disable the dc bias).

Figure 20 shows the B-H curve for the scrap transformer. This curve was interpreted from the B-H major loop on the oscilloscope display. The 1:1 transformer serving as the ac source in Figure 19 was replaced by a step-up transformer under the variac's control. This was required to drive the scrap transformer above its designed operating point for the purposes of plotting the B-H curve into saturation. Due to losses, the displayed flux density B was about 7% below the amount applied to the primary. The plot was adjusted upward by this amount.

Figure 20A shows how the normal magnetization curve plotted in Figure 20 can be generated. Note that the upper right cusp (+Hmax, +Bmax; points a,e,f,g) of each hysteresis loop is on the normal curve. Therefore, the normal magnetization curve is the locus of cusps of the gradually decreasing hysteresis loops. Use the oscilloscope method to display the hysteresis loop noting Bmax and Hmax for several different ac inputs. This procedure is done with an ungapped transformer.

To determine B and H from the oscilloscope display, it is required to know the number of turns in the windings. This could be learned

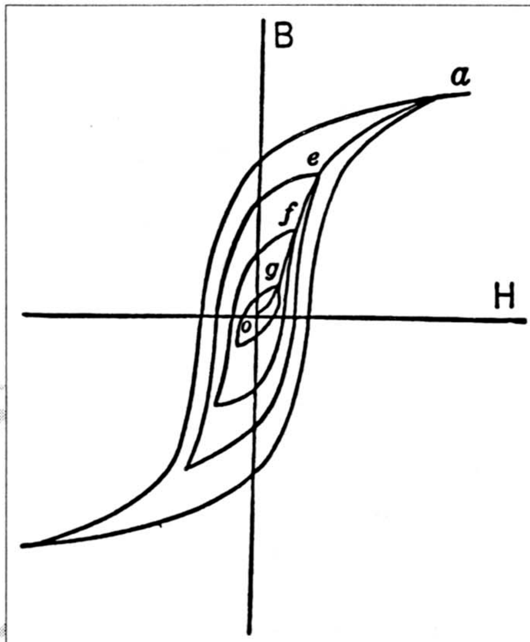

Fig 20A B-H curve generation

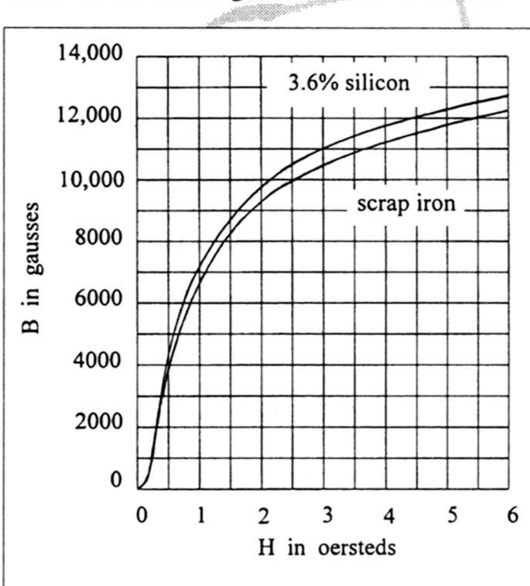

Fig. 20 B-H curves

by counting the turns when unwrapping the coil of the first scrap transformer. Then, before disassembling the second victim, connect it into the test setup of Figure 19. For you hacksaw types, there is a more clever way to find out the turns count, but only if the transformer has some extra window space available.

If the coil build doesn't completely fill the window, use the remaining space to temporarily add a 10 turn winding to the coil. Apply ac voltage to the normal transformer terminals and measure the voltage on the 10 turn winding. The ratio of the voltages is the ratio of the turns count in the windings. Since both voltages and one winding turns count (10) is known, the number of turns in the transformer winding can be calculated. For example, the scrap transformer's input terminals were connected to 120 Vrms, and 5.11 Vrms was measured across the 10 turn winding. Therefore:

Number of turns in the primary
= 10 x 120 / 5.11 = 235

This is valid if the 10 turn winding is added onto the existing coil. If the temporary winding is wrapped around an outside limb instead of the coil, it receives only half of the flux traveling in the E-I core, and the calculation is too high by a factor of two.

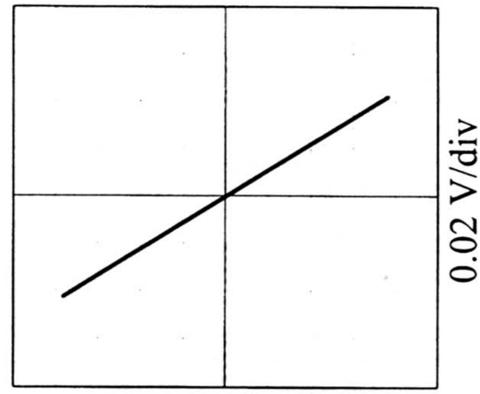

Fig. 21 SE OT B-H curve

Figure 20 also shows a standard B-H curve for 3.6% silicon iron. Although the curves were generated differently, they are similar which lends merit to the initial assumption that the scrap transformer's core material is silicon steel.

The RC network integrates the secondary's output voltage providing a signal to the oscilloscope Y channel that is proportional to the flux density B:

$$B = (Vy \times 10^8 \ R2 \ C) / (6.45 \ CSA \ NTs)$$

where:

Vy = oscilloscope trace voltage along the Y-axis

R2 = integrator circuit resistance

C = integrator circuit capacitance

CSA = cross-sectional area in square inches

NTs = number of turns in the secondary winding

The voltage across R1 driving the X channel is proportional to the magnetizing force H:

$$H = (Vx \ 0.4 \ pi \ NTp) / (R1 \ 2.54 \ MPL)$$

where:

Vx = oscilloscope trace voltage along the X-axis

NTp = number of turns in the primary

winding

R1 = series resistance in the primary circuit

MPL = magnetic path length in inches

The best accuracy is obtained when the integrator time constant is much greater (say 50 times) than the period of the applied signal:

R2 C > 50 x Period = 50 / Freq

If the input signal is 60 Hz and C = 0.22 uF, then R2 is greater than 3.9 Mohms. The values used in the integrator are not critical. Also, R1 should be non-inductive at the applied frequency.

A properly designed SE OT exhibits a straight-line minor hysteresis loop unlike the classic double-S shaped loop displayed in the textbooks. With the SE OT under test, sweep the bias current and notice the change (or lack of) in the oscilloscope display. The trace should remain a straight line for bias current values near the design value (90 mA) when the gap is properly sized. Raising the bias current higher will
result in curving of the oscilloscope trace.

I suggest using the minimum gap size that displays a straight line trace when the applied bias current is a little higher than the design value. Just how much higher is a matter of compromise between linearity and primary inductance (low end frequency response). I chose a value 10% higher than design.

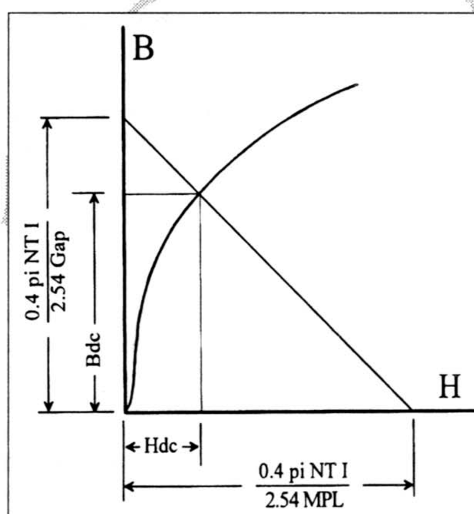

Fig 22 Graphical solution

The slope of the oscilloscope trace is the average permeability of the transformer core. The permeability is:

u = B / H

Using the equations for B and H above and a little algebra:

u = (Vy / Vx) (R1 R2 C MPL) / (CSA NTp NTs 3.19×10^{-8})

measured the B-H curve for the SE OT using the test setup in Figure 19 with R1 = 16.6 ohm, R2 = 3 Mohm, C = 0.22 uF, Vp = 200 Vrms at Ip = 90 mA. Figure 21 shows the uneventful looking oscilloscope trace. In this case, boring is good. The voltage on the Y-axis representing the flux density B was 0.086 Vpp and the X-axis representing the magnetizing force H was 0.76 Vpp. Therefore, the average permeability is:

u = 3487 Vy / Vx = 3487 x 0.086 / 0.76
= 394

The average permeability determines the inductance:

L = (3.19 x $(2736)^2$ x 394 x 3) / (9 x 10^8)
= 31.3 H

For a comparison, using the test setup in Figure 17 with R = 1242 ohm, Vtotal = 123.8 Vrms, and Vr = 13.28 Vrms:

Zt = 1242 x 123.8 / 13.28 = 11.58 kohm

XL = sqrt[$(11.58k)^2$ - $(136 + 1242)^{2]}$
= 11.5 kohm

L = XL / w = 11.5k / (2 x pi x 60) = 30.5 H

The two inductance measurements are in close agreement. Actually, the difference is surprisingly small considering all of the approximations and the oscilloscope accuracy.

Figure 22 shows how to graphically solve for the static flux density Bdc due to the dc bias current in the primary. Choosing Bdc and knowing the applied magnetizing force H determines the gap length. A typical value for Bdc is 8000 gausses. If Bdc is too high, the OT is likely to saturate on large amplitude, low frequency signals. Apply the method in Figure 22 to the B-H curve in Figure 20:

H = (0.4 x pi x 2736 x 0.090) / (2.54 x 9)

= 13.5 oersteds

The applied magnetizing force is off the scale, so extend the H-axis and draw a straight line from H = 13.5 through the B-H curve to the B-axis. Make the straight line intersect the B-H curve at the chosen Bdc value (8000 gausses). Read the B value where the straight line intersects the B-axis (9000 gausses):

B = (0.4 pi NT I) / (2.54 Gap) = 9000 gausses

Rearrange and solve for the gap length:

Gap = (0.4 x pi x 2736 x 0.090) / (2.54 x 9000) = 13.5 mils

Remember, the spacer thickness is half of the gap length. Spacer thickness is 6.75 mils

Note that the Magnequest and AudioNote transformers under test in Hodgson's article (reference 10) remain linear over a wide range. These SE OT's have lower impedance ratios and can achieve the same low end frequency response with less primary inductance. Also note that the core cross-sectional area is a little larger in the Magnequest than in the 5.1 kohm example discussed here. That combination of low impedance and a large core makes for a very solid design. This suggests that the 5.1 kohm 90 mA SE OT needs a larger (possibly exponentially larger) core to achieve the same headroom offered by the lower impedance SE OT.

Adjusting the Gap

The best way to adjust the gap, particularly when using an unknown core material, is by the oscilloscope method. As an initial guess, use the spacer thickness determined by the graphical solution. If an oscilloscope is not available, use the test setup in Figure 17 to find the correct gap length. Select the gap that exhibits constant inductance as shown in Figure 18. For the graphical solution, use the standard B-H curve for silicon steel to determine the initial guess for the spacer thickness.

Measurement Summary

The 5.1 kohm SE OT performs well. Driven by an 845, the response is flat through the midband and rolls off smoothly at each end. The lower -3dB point is around 8 Hz and the upper -3dB point occurs at 33 kHz.

The OT will easily deliver the design goal of 10 watts across the full frequency range. As expected, it is in the low frequencies that the power output is restricted. There is a limit to the peak flux density that the core can linearly sustain. Since the flux density is inversely proportional to the signal frequency, the demands on the OT are greater as the bass goes deeper. The output begins to falter (onset of waveform deformation due to core saturation) at the extreme low end and is probably limited to 12 useful watts at 20 Hz.

The secondary can be configured for a 2 ohm load. Connecting a 4 ohm speaker load to the 2 ohm output tap reflects 10 kohm into the primary. To get the same output power requires a larger signal from output tube, and consequently, a greater voltage swing from the driver circuit. The larger voltage signal appearing on the primary raises the core flux density pushing the core towards saturation sooner. Raising the transformer impedance for less distortion in the output tube can be more than offset by an increase in distortion in the OT and the driver circuit.

The secondary can be configured for a 16 ohm load also. Connecting an 8 ohm load to the 16 ohm tap reflects 2.5 kohm into the primary. There is plenty of inductance for good low end performance when driven by a 2A3. Because the bias current is lower, the gap could be squeezed for even less phase shift. The down side is that the ohmic resistance of the windings have grown to a larger percentage thereby increasing the insertion loss. Once again, if a different transformer impedance is desired, the best results come from designing for it in the first place.

In either case, whether matching up or down, it still might be worth a try. Don't be discouraged just because the math says so. Just be aware of the limitations.

A Better Model

The simple transformer model used thus far predicted a higher rolloff frequency (Fh) than actually measured. Looking back at the rolloff equations, it can be seen that the high frequency rolloff is a function of not only parasitic capacitance and leakage inductance, but also the reflected load resistance and the tube's plate resistance. Interestingly, the relationship of resistance to Fh is different if the main cause of the rolloff is due to capacitance than if it is mainly due to leakage inductance. When the rolloff is due chiefly to inductance, Fh occurs higher as the resistance increases. Whereas, when the rolloff is due mainly to capacitance, Fh moves lower as the resistance increases.

Increasing the resistance can be accomplished by either changing to a tube possessing a higher plate resistance (812, 10, etc.) or by changing the load resistor connected to the 8 ohm tap. As an experiment, I investigated both changes. I measured the Fh of the OT when various loads (4, 6, 8, 16, 100 ohms) were connected to the 8 ohm tap of the OT being driven by a 2A3 (800 ohms), 845 (1700 ohms), 812 (4600 ohms), and 810 (10,000 ohms). Because Fh moved lower as the resistance was increased, it was determined that parasitic capacitance was the dominant cause of the rolloff.

It is unexpected that the chief culprit is capacitance. The primary's self capacitance is well below the amount required to rolloff the OT at 33 kHz. After all, it was the leakage inductance that forced the coil to be interleaved in the first place.

The missing pieces in the puzzle are stray capacitance and the bridging capacitance. Stray capacitance appears from each winding to the core (ground). The bridging capacitance occurs at the interface where the primary winding meets the secondary winding. Stray capacitance adds directly to the primary winding's self-capacitance and similarly in the secondary. The bridging capacitance affects the primary, the secondary, and also produces a resonance effect (frequency dip) with the leakage inductance.

In this design, there are eight interfaces; each is 10 mils thick. The interface capacitance can be calculated based on the coil geometry as before:

Co = 0.225 DC ALT Wb / d (in pF)

where:

DC = the dielectric constant of the insulation

ALT = the average length of a turn in inches

Wb = the width of the bobbin in inches

d = the thickness of the insulation in inches

The total static bridging capacitance consists of these eight interface capacitors in parallel. This total capacitance can be measured directly with an impedance bridge. The effect of this capacitance on the rolloff though is different from its static value. The effective value depends on the location of the interface within the coil and the applied differential ac voltage at that interface.

Figure 23 shows two adjacent windings forming a parasitic capacitor. The static value of this capacitor can be calculated or measured. The effective value of this capacitance is derived from the following as depicted in Figure 23:

$$C_{eff} = C_o \times (V_{ab}^2 + V_{ab} \times V_{cd} + V_{cd}^2) / 3V^2$$

The derivation will be omitted here (see reference 4). The important thing to note is that the effect is a function of the differential voltage squared. In order to minimize the effect of the geometric capacitor, it is imperative to minimize the voltage gradient. This also explains why winding polarity and Z-winding versus U-winding affects the frequency response.

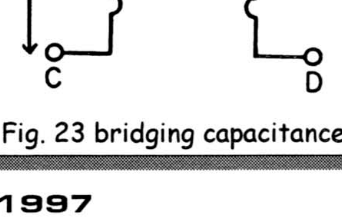

Fig. 23 bridging capacitance

Figure 24 shows the improved model. Cp and Cs are the self-capacitance of the primary and secondary windings as previously described. Cps and Css are the stray capacitances of the primary and secondary windings, respectively. Note that because this is a high ratio step-down transformer, the secondary capacitance C2 can be omitted with little loss in accuracy.

C1 = Cp + Cb + Cps/3

$C2 = [Cs + Css/3] / N^2$

C3 = Cb / N

The tricky part lies in the interpretation of Cb due to the coil interleaving. A multiplier, Cbf, is used to account for the coil geometry's effect on the bridging capacitance. Examination of the coil layout determines Cbf. The interface capacitance Co is multiplied by Cbf giving the bridging capacitance Cb as used in the model.

Refer to Crowhurst's article on capacitance (reference 6) for a table of Cbf based on various coil interleaves. Recognize that from the primary winding's point of view, there is little difference between ground, the core, a shield (if used), and the secondary winding - they are all at low potential. The small difference between ground and the secondary winding can be accounted for by a turns-factor multiplier. Because N is much greater than one (25 >>1), the multiplier is nearly unity, and therefore omitted from this model. Introducing a grounded shield between the primary and secondary eliminates the bridging capacitance (but adds capacitance to ground). This shield makes the transformer rolloff frequency independent of the secondary winding's polarity, and also interrupts the Cb-Llk resonance removing the amplitude dip at the resonant frequency.

I have built a 1:1 interstage transformer without shielding between the primary and secondary. In this case, the turns-factor multiplier can not be omitted. Using a non-inverting connection, the bandwidth extends pass 200 kHz. When connected in opposite polarity, the interstage transformer begins to roll off below 400 Hz.

Table 2 shows the predicted response of various coil interleaves. This is based on the OT designed in this article as driven by an 845 and connected to a 7.4 ohm load. The "Type" field describes the interleave pattern. The designed OT is of the "5p4s" type - it has 4 equal secondaries surrounded by a primary divided into 5 unequal sections. The unequal sections always follow the same interleave ratio: the first and last sections have one-half the number of turns contained in each of the innermost sections.

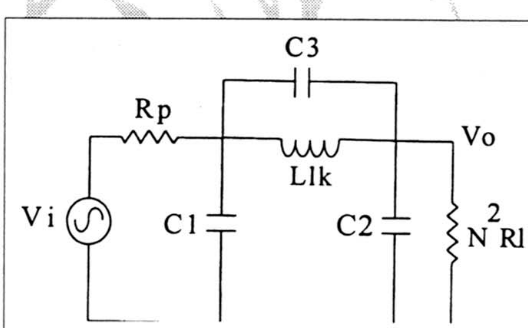

Fig. 24 High frequency model

The "Ti" field is the thickness of the insulation in mils used at each primary to secondary interface. The "C1" field is the effective value of all capacitance reflected into the primary winding assuming a dielectric constant of 2.8. The "Llk" column contains the leakage inductance in milli-henries. "Fh" is the predicted -3dB upper rolloff frequency in kilohertz.

The first eight entries are identical except for the interleave pattern. Based on Fh, entries 1 and 2 are unacceptable, entries 7 and 8 are marginal, and entries 3 through 6 show promise. Entry 7 is the model prediction for the designed 5.1 kohm OT which has a measured Fh of 33 kHz. Recalculating with the measured values for C1 and Llk, the model accurately predicts an Fh equal to 33 kHz.

The window size is fixed so the coil build is also fixed. Choosing a less complicated interleave pattern requires fewer interfaces to be filled with insulation. This means that thicker insulation can be used in each interface and still maintain the same coil build. This thickening lowers the bridging capacitance at the expense of increased leakage inductance. Entries 9 through 10 show this effect. The Fh value for entry 9 does not improve over entry 1. In this case, the leakage inductance is dominant in determining Fh, so increasing the insulation only makes matters worse. Entries 10 and 11 show that the right combination of interleave pattern and insulation thickness extends the frequency response.

Comparing the "3s2p" versus the "3p2s" shows the effect of swapping the locations of primary and secondary. The factor Cbf changes slightly and the stray capacitance can be ignored in the "3s2p" case. The stray capacitance forms between the winding and the core. Only the innermost winding layers and the outermost winding layers are close enough to the core to contribute. The "3p2s" has primary stray capacitance but virtually no secondary stray capacitance. The "3s2p" is just the opposite. But because this is a high-ratio step-down transformer, the secondary stray capacitance can be safely ignored.

Calculating the stray capacitance is difficult. The innermost winding layer is separated from the tongue of the core by the thickness of the bobbin. As usual, the geometric capacitance Co can be calculated provided that the thickness and dielectric constant of the bobbin is known. Use the actual length of a turn at the inner-most layer instead of the average length of a turn (ALT). The outermost winding layer is separated from the limb of the core by the final insulating wrap on the coil and whatever air space (margin) remains in the window. If the coil build completely fills the window, then very little air space remains, and the stray capacitance will be larger. Unlike the previous calculations of Co, the full length of a turn is not used here. The full surface area of the coil does not contribute significantly, only the area within the window next to the core limb.

Good coil design maximizes the use of the core window. It is important that the coil build does not exceed the height of the window or the transformer can not be assembled. A margin is allowed for to ensure that the coil will just fit. Coil designs that have the primary (high voltage) winding in the outermost layer must allow extra margin to prevent arc over to the core limb. The final wrap insulation must be at least as thick as the insulation used at each primary to secondary interface. A coil design with the secondary (low voltage) winding in the outermost layer does not suffer this extra constraint.

Most often, the stray capacitance is ignored. It tends to be small, hard to accurately calculate, and only one-third of its static value effectively contributes to the capacitance in the primary winding. When the secondary winding is the outermost layer, the effective value is insignificant. For comparison, here are the effective values in the 5.1 kohm OT: primary winding self-capacitance is 0.69 nF, bridging capacitance reflected to the primary is 3.05 nF, primary winding stray capacitance is 0.086 nF. The point is that whether it is modeled or not, stray capacitance does exist, and its small influence differs for different winding layouts.

	Type	Ti	C1	Llk	Fh
1	2p1s	10	1.44	52.8	20.3
2	2s1p	10	1.97	52.8	19.3
3	3p2s	10	2.41	14.5	41.2
4	3s2p	10	2.62	14.5	39.0
5	4p3s	10	3.30	7.1	38.0
6	4s3p	10	3.41	7.1	36.9
7	5p4s	10	4.18	4.3	30.4
8	5s4p	10	4.39	4.3	29.0
9	2p1s	40	0.96	69.0	16.1
10	3p2s	20	1.60	17.2	47.4
11	4p3s	13	2.72	7.6	45.0
12	3p2s	17	1.78	14.8	49.1
13	3s2p	17	1.86	14.8	47.7

Table 2 model predictions

For comparison, I chose to wind another OT utilizing a different interleave pattern. The number of primary layers and the secondary layout (double layer of single/bifilar combination) were to remain unchanged in order to keep the comparison valid. This ruled out the "4p3s" type because the secondary windings are incompatible with the "5p4s". I chose the "3s2p" type (entry 13), but only increased the interface insulation thickness to 17 mils instead of the full 20 mils because the original coil build was such a tight fit. For equal splitting, the number of primary layers was changed from 19 to 18. The coil layout is shown in Figure 25. Notice the symmetry of the secondary layout. Always arrange the windings for best symmetry if possible.

Entry 10 can not be fairly compared against entry 12. The results would indicate that 20 mils was too thick, and that 17 mils is a better choice. It is true that above a certain thickness, Fh begins to fall as the leakage inductance takes over. But that is not the case here. Entry 12 has fewer turns in the primary which lowered Llk and improved Fh. Remember that fewer turns will affect the low frequency response in two ways. First, it raises the -3dB lower rolloff point due to less inductance. Secondly, it lowers the maximum power available because it raises the flux density in the core. The core will now saturate at a smaller applied voltage swing. There is not enough core cross-sectional area to support the increased flux density brought on by fewer turns.

The results shown in Table 2 can not necessarily be applied to other tube and OT combinations. The "3p2s" layout may not be the best in all cases. The values particular to a different application should be inserted into the model for analysis. The same general trend should emerge indicating the best tradeoff between coil interleave complexity and interface insulation thickness.

The new SE OT depicted in Figure 25 has an Fh of 52.5 kHz. This is much higher than in the first design and slightly better than predicted. Recalculating with the measured values for C1 and Llk, the model predicts an Fh of 48.7 kHz.

Model Limitations

The transformer model still has limitations. The model described in this article ignores winding resistances, winding polarity, secondary capacitance, and dissipation factors of all the capacitance. Furthermore, it uses lumped values instead of distributed ones. Increasing frequencies with long wire lengths indicates that transmission theory may be required for accurate modeling. The lumped values used are the averages across the entire winding. The actual transformer consists of multiple windings that each behave a little differently from the average. Better results are obtained by considering each interface separately and then combining the responses vectorialy. This may be most evident when using the 16 ohm secondary taps. In this case, some of the secondaries are in series instead of in parallel and their individual responses add differently.

I've written a computer model to take as many of these variables into account. It differs only slightly as long the simpler model is not asked to predict extreme cases - a 2A3 connected to a 5.1 kohm OT wired for a 4 ohm load but actually connected to 16 ohms at 100 kHz. I have tested the model's predicted results against actual measurements in many tube, load, and output configurations. I have even soldered small capacitors into the windings to simulate a change in parasitic capacitance. The transformer model's prediction of Fh is suspect above 50 kHz. At this point, small differences in the parasitics begin to have large effects on Fh. What remains consistent and valuable, is the ability to compare different layouts. The model may not predict the actual value of Fh, but it reliably predicts relative values of Fh. The user can determine which layout will have the greater bandwidth.

One particular area which requires a closer look concerns the multiplier Cbf as it is applied to the bridging capacitance. The table in Crowhurst's article is based on an average value of the interface capacitance. When the insulation thickness is the same at each interface, the static capacitance is different at each

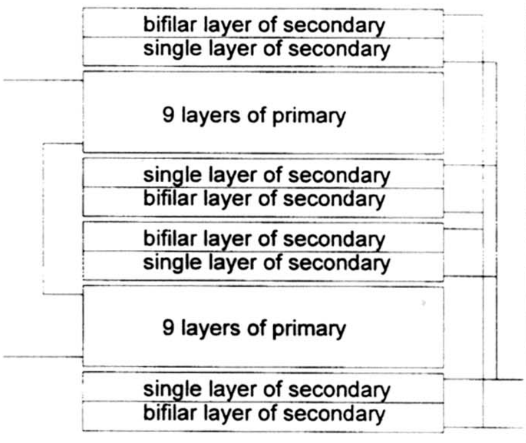

Fig 25 3s2p coil layout

location because the winding length is different. For the original OT, the winding length at the outer-most interface is twice that at the inner-most interface, which means that the bridging capacitance at the outer-most interface is twice as large as that found at the inner-most interface. For best accuracy, Cbf should be broken down and applied individually at each winding. Using the values as supplied in the table should be sufficient in most cases.

It seems logical to compensate for the different winding lengths by juggling the insulation thickness accordingly. The designer could add a little here and take out a little there and keep the total (coil build) constant. The distributed bridging capacitance is equalized, but the leakage inductance also needs consideration. Do not reduce the interface insulation thickness too much. Although the ac voltage gradient may be small at a particular interface, the dc voltage differential must still be supported by the insulation thickness.

It may not seem worthwhile to go to some of these lengths in the design process. After all, if the OT rolls off above 20 kHz, what does it matter? Well, an OT with a well behaved rolloff is better than one with a series of high amplitude resonances. It is still being argued as to what is audible, and how high frequencies intermodulate into the passband. The application of global feedback requires a well behaved response or you are likely to end up with a high frequency oscillator. Furthermore, I'm all for the elimination of capacitors in the signal path. The parasitic capacitors in the OT are not exactly of the audiophile variety. Their insulation is composed of the coatings on the wire and whatever the person winding the OT used as layer insulation.

Comparison and Conclusion

Assuming that the reader is now armed with the knowledge of how to design and build his own transformer, how will his effort compare with the commercial offerings? What does Mike LeFevre of MagneQuest know that you don't? What can the manufacturers do that the home builder can not?

What it comes down to is design versus materials used in the design versus implementation of that design versus cost. There is no magic. If you have the best design, using the best materials, constructed of the highest quality, you win. The reality is two fold. What is really the "best", and at what cost? Many engineering decisions determine the performance of the finished product. There is not a special parameter written on a spec sheet that can be used alone to judge one transformer as better than its competition. Because one transformer has a more complex interleave structure does not make it better. Because one OT specs a higher bias current does not make it better. One must consider the total package when choosing.

There are many questions that I would not expect an OT manufacturer to answer. Partly this is due to nondisclosure of their "secret recipe", and partly because that answer may not stand alone for a basis of comparison and requires a lot of explanation. I think that "What kind of insulation do you use?" falls into the first category and "What is your interleave factor?" goes in the second category. While I may not expect exact answers, I would expect a manufacturer to be able to knowledgeably discuss the engineering decision in general and explain the reasoning behind the choice. Because of these missing answers, it can be hard to accurately judge the home builder's OT against the commercial offerings.

One thing in the commercial world's favor, is their ability to obtain high quality iron for the core material. This assumes that they know which is the better alloy and that it actually makes a difference in the particular application. One thing is certain, some irons can support a higher flux density than others. This could allow for a greater range of useful linearity and lower losses in the iron core. There are higher grades of laminations available than what is likely to be found as surplus as I used for my OT. It may be possible to buy high grade laminations from a transformer manufacturer or possibly have them special ordered for you. Even when limited to only the standard grades, all is not lost. I recently learned that the iron used in the OT for the WE91 amplifier is similar to that found in power supply transformers.

A high quality commercial OT undoubtedly uses better insulation than what I can find in my kitchen drawer. I would expect this to be top secret. Unless one of you readers out there can help out, it looks like the commercial guys have an edge there. As much as I'd like to know the answer, I am not requesting anyone to divulge anything told in confidence.

The subject of potting, impregnation, or encapsulation has not yet been discussed. I don't know what is best, but I am firmly against any rigid, unyielding, high dielectric constant potting compounds being impregnated into an OT's windings. The windings and the insulation must be protected from the elements otherwise the performance will deteriorate over time. All impregnants that I have read about raise the effective dielectric constant of the insulation. This of course raises the parasitic capacitance which so much work has gone into eliminating. As it turns out, the designer is dealing with the lesser of two evils. Impregnation will raise the capacitance noticeably, but will do so in a predictable and stable (over time) way. Left to the elements, the capacitance will eventually rise to an even higher value that changes with the weather.

I imagine that a good encapsulation substance has a low dielectric constant and is firm but not unyielding. I would hope that its consistency would tend to damp out vibrations that cause an OT to sing with the music. In my other hobby, I design, grind, and polish my own optics for use in a home made telescope. The optics require special coatings that I can not apply myself. There are optics businesses that will apply the coatings for a reasonable fee. Perhaps an OT manufacturer can be persuaded to offer a similar service. The home builder could get his OT professionally encapsulated in a high quality compound

(Continued on page 20)

ones and two NOS UX-250s, thank you very much. No, they ARE NOT for sale!

So I do my usual corner cutting, grab a 5842 preamp circuit I've been dinkin' with and a stack of rack power supplies, and hack together a circuit using a single 5842 cap coupled to the grids of paralleled 50's, which are loaded by a 30H or so choke, and coupled to the nickel critter by my gonzo composite cap.

After some serious tweaking to get the 5842 stage so it would drive into the 5K grid resistor of the paralleled 50s, this amp was ready for bear.

It took on Gary Dahl's latest 5842/Tango NC18 interstage/ Vaic VV30 / Magnequest FS-030 creation, and while it lacked a bit of top end, and perhaps some bass extension in comparison, it easily matched or beat the Vaic amp in midrange smoothness and speed, with a definite edge in the 'twinkle'.

I ran some specs for grins and here's what I got:

A single 50 is spec'd for 4.8 watts into 4800 ohms at the operating point I used, so we are definitely on the money with our 10W 5% figure. And look how gracefully this thing overloads. It's a keeper.

power	THD	3rd harmonic
1 watt	0.58%	45dB down
4.7 watts	2.0%	36 dB down
10 watts	5.0%	40 dB down
14 watts	10.0%	35 dB down

Lesson 3:

You can squeeze a little more out of the same size core with parallel feed, but don't push it if you use nickel.

Oops, out of room. Next month the 50 schematic, and results of my trials with an interstage coupled 5842 driver stage on the 50 amp, and Mikey's new 50H audio chokes as plate loads on the Afterglow para feed project.

may meeting

Sunday, May 4, 12 noon. Bring your latest project, and your favorite vinyl rocording.

(Continued from page 15)

without the manufacturer ever letting on to just what the secret substance is.

I think that the home builder can potentially design as well as the commercial guys. The information is available, computers are ready to make the task less daunting, and everyone is limited by the same rules of physics. I know that at least one highly regarded manufacturer has an archive of designs put to paper by some of the best transformers experts of their time. These may be hard to beat, but materials have improved, and I just hate to think that best there is has already been done.

It will take a lot of practice for the home builder to equal the craftsmanship of the best built OT's available. I sometimes think that good winding technique is part art form. I have removed an entire layer on more than one occasion because it just didn't turn out right. The good news is that the home builder is not subject to any undo time pressures or schedule crunch. You should be able to take as much time as necessary to get it right.

The original intent of this article was to share what I have learned about designing and building a single-ended output transformer. I wished to equip the reader with the knowledge required to build a transformer of his own design. I hoped that the reader would subsequently design his own OT and report back on how it turned out. Anyone should feel free to copy the 5.1 kohm SE OT exactly as described here for their own use if so desired. It is a fine design as is the second one, but can still be improved upon with only minor changes.

Although this article covers a lot of ground, it is not a cookbook. I do intend to write a "How To Design And Build An Audio Transformer" style of cookbook with all of the step by step instructions spelled out. The book will include finished designs of output transformers for the more common audio tubes (2A3, 300B, 845, 211), cathode negative feedback OT's for transmitter tubes (812, 805, 810), and interstage transformers. All of the theory, modeling, and calculations will be included. The transformer modeling software will also be available. The addition of power supply transformers and filter chokes is under consideration.

JF