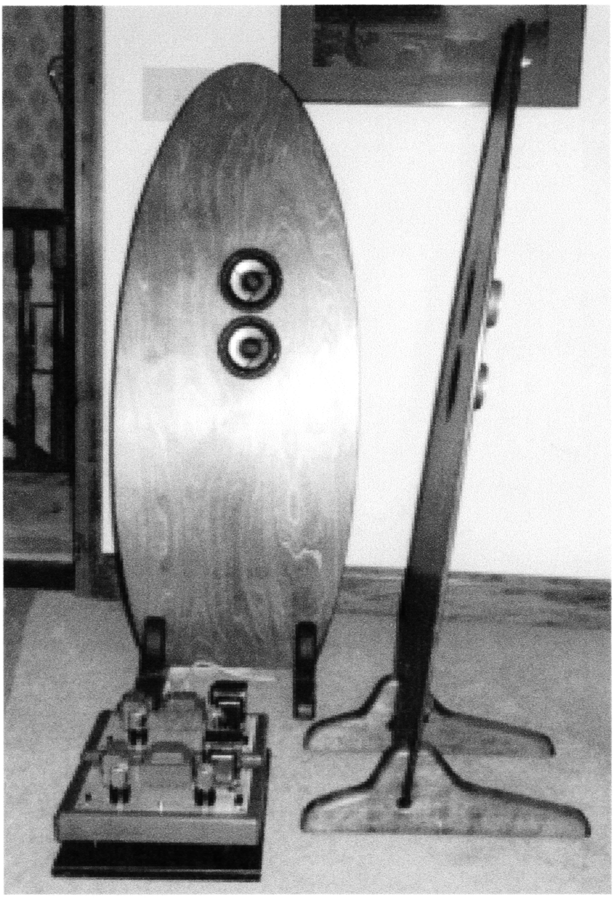
VALVE

in this issue:
Vacuum State of the Art Conference
winding a SE output trans - conclusion
what's all this about parallel feed?



Ray Kuehlthau's S.E.X. system

volume 4, number 4 April 1997

VALVE

the monthly magazine
for tube audio eXperimenters

Editor and Publisher
Dan "Dr. Bottlehead" Schmalle
Chief Administrator
Eileen "In Charge" Schmalle
Contributing Editors
David "Full Track" Dintenfass
Doug "Lazarus" Grove
Resident Smart Guy
Paul "Braniac" Joppa
Mapmaker
George "Always" Wright
Big gun OEM advisors
John "Smoothplate" Tucker
Michael "Airgap" Lefevre

Our mailing address:
VALVE
P.O. Box 2786
Poulsbo, WA 98370
by phone: 360-697-1936
(hours: 9-6 PST, Mon -Fri)
fax: 360-697-3348
NEW e-mail - Bottlehead@prodigy.net

Rates:
Membership/Subscriptions:
$25.00 per year (12 issues)
Foreign Subscriptions:
$35.00 per year (12 issues)
Please make checks payable to VALVE
MC/VISA/AMEX/DISC accepted

editor's thing

Yo, we did it!

The first VALVE magazine Vacuum State of the Art Conference will be held at the Silverdale on the Bay Resort Hotel on Labor Day weekend, August 30, 31 and Sept 1.

We have tentatively reserved one wing of the hotel, about 65 rooms. We hope to fill 20-30 of these with exhibitors demo systems and product displays. The rest of the rooms will be available to attendees, and there are several other motels within a reasonable distance.

We also plan on having two seminar rooms going. Paul Joppa will be our Seminar Coordinator. He is looking for help from someone who could make calls and line up some speakers for the 20 minute presenatations we will have at the beginning of each hour long session. The rest of the session will be a question and answer, argue the fine points, informal kind of format so everyone gets a chance to deal directly with their favorite guru.

We will have a room dedicated to a display of fine vintage audio gear (call me if you have something really special you could bring), and another room we will call the "15 minutes of fame room" where bottleheads can bring their latest DIY creation and demo it in a reference system.

In conjunction with this room we will have awards for the best sounding gear and the most beautifully crafted gear.

Mike Lefevre will be awarding a pair of transformers to the lucky first place winner in the craftsmanship category, so get those sawblades sharp and give your copper merchant a call.

Nuts About Hi Fi is a few minutes walk from the hotel, and I'm certain Bill Benson will have something special cooking for the show. We have a few ideas for the Sakuma concert venue, and we will fill you in on where this will occur too.

David Robinson and Lynn Olson of *Positive Feedback* magazine have offered their assistance, as has *Sound Practices'* Joe Roberts (thanks much, guys), and we hope to have other members of the press present too.

We will try to make the exhibitor room fees as reasonable as possible. I will have a better handle on this soon, and we will be sending out applications for exhibitors later in the month.

Bottlehead Rick Francis has volunteeered to post a web page which will give the latest info on events, logistics and prices as it develops.

Following the show, we will be giving a master amplifier design and build class. John Tucker and I are still developing this class, and we will have more info on it soon.

Picture sitting in a lab type environment, discussing basics as well as the finer points of tube audio design for the first part of each day, and then going hands on, building your design (6DN7, 2A3, 300B, direct coupled, RC coupled, etc. etc.) onto a basic, universal type of chassis all afternoon, while a top dog reference system accompanies your inspired creativity as you sit at your personal workbench. Before the class ends you will have a chance to run your finished product on the reference system and get a frequency response curve and THD readout to take home with your amp.

You can call the hotel for room reservations at 360-698-1000.

Doc B.

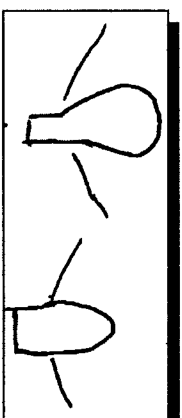
the cover

Is that a cool system or what? Naturally, Ray has changed his S.E.X. speaks since this photo was taken. We talked about removing the dust cover from one driver, as discussed a while ago in Letters, and Ray gave it a shot:

"After slicing off the dust covers I dabbed some paint remover on the remaining glue ring and lifted it off with the corner of a credit card. No sign that a dust cover ever existed.

"Then I turned up a phase plug (Ray builds race cars for a living) out of 1-1/4" diameter nylon. I copied the Lowther-Voigt shape out of an old *Sound Practices*.

"Now I'm thinking that shape is more correct for whizzer cone technology. I will try a reverse parabola next and compare."



did you just tune in? here's what's happened so far...

Back Issues

Volume 1 - 1994 issues - $20

a Williamson amp; Dyna Stereo 70 mod bake-off; converting the Stereo 70 to 6GH8's; a QUAD system; triode input Dyna MkIII; MkIII vertical tasting; smoothing impedance curves; Altec A7; Ampexes Nagras and ribbon mikes; Triophoni, a 6CK4 amp; audio at the 1939 World's Fair; books for collectors and builders; V.T. vs. R.M.A. cross reference; FM tuner tube substitutions; Big Mac attack - the MI200; 6L6 shootout; a vintage "audessey"; more FM tuner mods; vintage radio mods; Heathkit rectifiers; PAS heater mod.

Volume 2 - 1995 issues - $20

Rectifier shootout, tube vs. solid; FM 1000 recap and meters; single ended 10 amp; triode output W-4; Optimus 990 - speaker for SE?; star grounds; tuner shootout; Living Stereo, vinyl or CD?; World Audio SE integrated; firin' up - smoke checking; Brook 12A schematic; 6C33 vs. 3C33; Heathkit power transformers; 6B4's + Magnequest = SEcstasy; W5 mods; triode operating points; Dyna restorations; Marantz 7,8 and Scott LK150 impressions; hackable vintage gear; Quasimodo - PP 805 amp; restoring a Scott 340 in 75 minutes; a dream system for 78's; cartridges and styli for 78's; Restoring a Lowther, Part 1&2; easy tube CD output hack; 6ER5 phono preamp; 304TL & 450TH SE operating points; hypothetical DC ESL amps.

Volume 3 - 1996 ($25):

Single Watt, Single Tube, Single Ended, an amp for Lowthers; the Vintage Speaker Shootout of 1996, QUAD vs. Lowther, vs. A7; the Voigt Loudspeaker, the Single Ended eXperimenter's kit; cathode coupled SE 6AS7 amp; how to build the Superwhamodyne; refoaming AR woofers; mesh plate tubes; rebuilding QUADS; QUAD amp filter surgery; single gain stage amps; the Brooklet, and Brookson, choke loaded PP 6080 amps; transformer coupled PP 6DN7 amp; the Iron Maiden; Building the Lowther Club Medallion; the TQWT, a tapered pipe enclosure; IT 300B amp.

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 gapneeds 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.

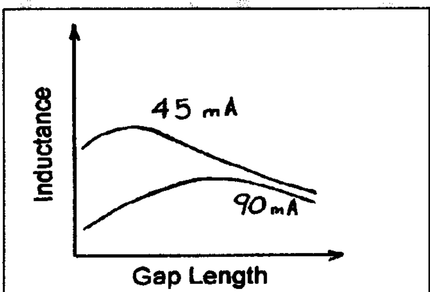


Fig.16 Inductance vs. Gap Length

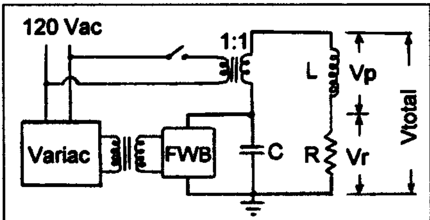


Fig. 17 Inductance measurement circuit

This is the gap size suggested by the Hanna curve.

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 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.

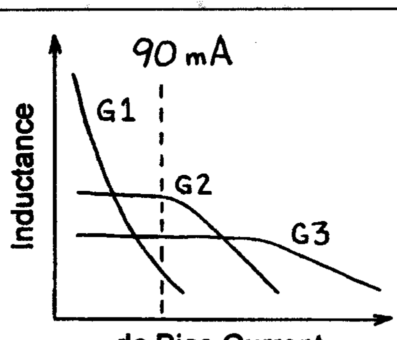


Fig. 18 Inductance vs. Bias

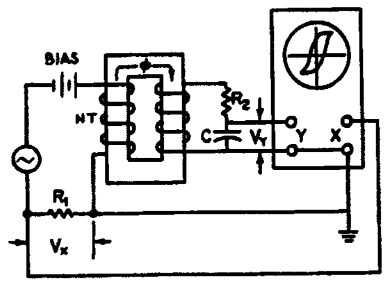


Fig. 19 B-H curve display circuit

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.

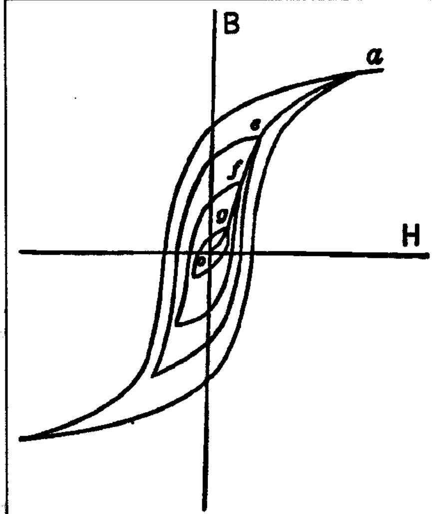


Fig 20A B-H curve generation

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 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:

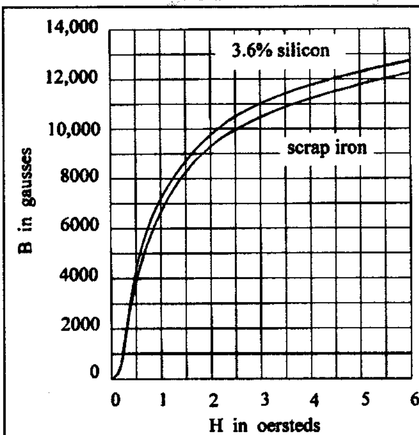


Fig. 20 B-H curves

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.

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:

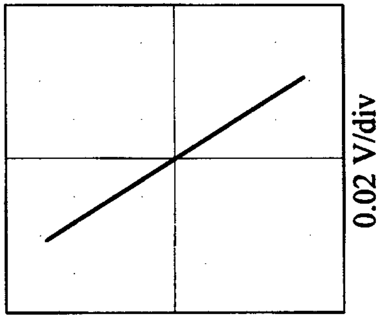


Fig. 21 SE OT B-H curve

$B = (V_y \times 10^8 \; R2 \; C) / (6.45 \; CSA \; NT_s)$

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:

$H = (V_x \; 0.4 \pi \; NT_p) / (R1 \; 2.54 \; MPL)$

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:

$R2\,C > 50 \times \text{Period} = 50 / \text{Freq}$

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.

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 x 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

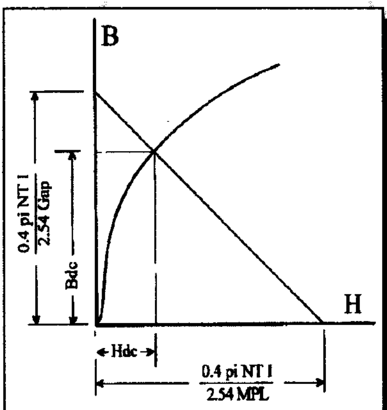


Fig 22 Graphical solution

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 OTs 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.

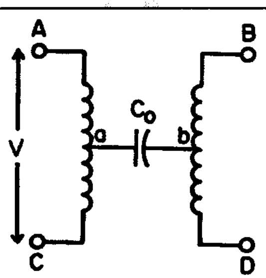


Fig. 23 bridging capacitance

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:

Ceff = Co x (Vab^2 + Vab x Vcd + Vcd^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.

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.

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.

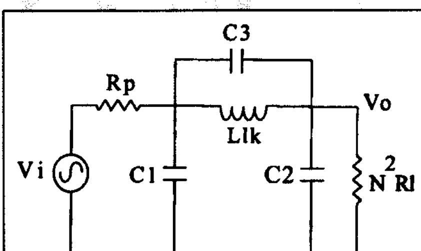


Fig. 24 High frequency model

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.

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.

	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

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 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

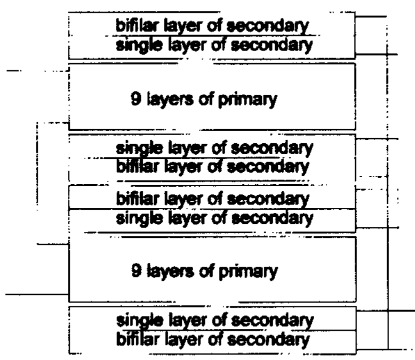


Fig 25 3s2p coil layout

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 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.

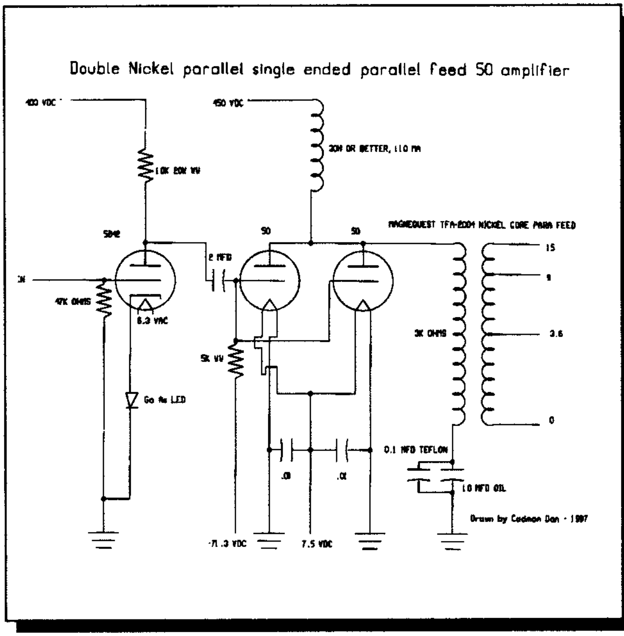
The original intent of this article was to share what I have learned about designing and building a single-ended output transformer. I wished to equip the reader with the knowledge required to build a transformer of his own design. I hoped that the reader would subsequently design his own OT and report 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.

(Continued on page 20)

what's all this about parallel feed?

by Doc B.



It all started with one of my weekly phone calls from Mike Lefevre.

"Hey Schmoofy, I got an idea I been working on for about a year and a half. Nobody else will touch it, and I know you'll try anything that doesn't blow out your microwave. You want a crack at it?"

Cheap stunt. Mikey knows I'm a sucker for something nobody else will touch.

A few weeks later I get two pairs of TFA-204 output trannies, only these don't look like ordinary SE output trannies. They are interleaved, no air gap. One pair looks like it has good ol' M6 steel lams, and the other has the dull sheen of nickel. Hmmmmm.

In the meantime Mike has sent me some schematics from Costriure Hi-Fi. These schemos show caps coupling the output trannies to the plates of the output tubes, and a choke connecting the plate of the tube to the B+ supply.

"Parallel feed (also called shunt feed)", says Mike.

"The cap blocks current from flowing through the transformer. No DC in the trannie core to saturate it. Hence no need for air gapping, and you can use nickel, which is normally hypersensitive to DC."

"Gad", you purists say, "You're putting another one of those poisonous capacitor thingies in the signal path. What's wrong with you? That can't possibly sound good."

Ha. Double Ha.

Give this some thought.

In a typical class A SE power amp, the output transformer floats between the plate of the output tube and the power supply. Think about how the AC signal has to go to find a low impedance path to that output trannie from the ground at the center tap of the power transformer.

Lessee, through the power trannie, through some diodes, through some of those poisonous capacitor thingies, and great big slow ones at that, through a choke, maybe through a resistor or two.

Now what does it have to go through to get to a parallel feed transformer?

Wire. Direct to ground. Now that's low impedance!

Yes there's a coupling cap in the mix now. But who says we have to put it between the output transformer primary and the plate of the output tube?

Can we take it out of the direct signal path between the plate and the primary, and instead put it in the 'to ground' leg?

Enter Sandy Ong, a closet parallel feeder (although he prefers the term shunt feed) for quite some time.

Mike asked Sandy to call me, and we both got excited that the other guy was thinking this parallel fed stuff was special. Sandy has an 811 based Class A2 parallel feed SE amp that we will discuss in detail in a future issue. He was so tired of hearing "that can't work" that he wouldn't even tell folks it was parallel feed when he auditioned it for them.

I asked Sandy about moving the cap to the ground side of the trannie, and he said it works better, making the image wider from his amps.

So I put this parallel feed deal into a couple of

Introducing Brooklyn. Push-pull transformers that sound single-ended.

Recent developments have led us to an even better sound! If you've been waiting to try a Brooklyn transformer, wait no longer. This month we are incorporating improvements in the core construction that give even better bass response, making Brooklyn transformers the best entry level push pull transformers going.

PART #	PRIMARY IMPEDANCE	POWER LEVEL	MAX. PRIMARY DCMA PER SIDE	DCMA UNBAL	RETAIL EACH
B14	12,000 CT	10W	40	4	$100
B15	10,000 CT	20W	50	5	$125
B17	9,000 CT	30W	50	5	$140
B18	8,000 CT	15W	45	5	$120
B20	6,600 CT	30W	70	7	$140
B21	5,000 CT	20W	80	8	$120
B23	4,000 CT	50W	100	10	$150
B24	3,000 CT	15W	75	7.5	$125
B27	1,500 CT	30W	150	15	$135

Recent measurements reveal the B24 response to be -.35 dB @ 10Hz and -.8 dB @ 100kHz!

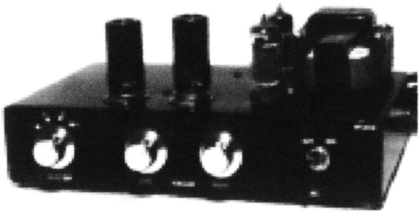
Note: Above units available with Ultralinear taps for an additional $6.00 per unit
Secondary impedances far all units are 2,4,8,12 & 16 ohms.
Guaranteed minimum frequency response is +/- 1 dB, 30Hz to 20 kHz.
All units supplied with vertical bell end caps.
All prices herein are special introductory - prices subject to change without notice.

Brooklyn

Brooklyn, P.O. Box 967
Cherryville, PA 18033
(215) 288-4816

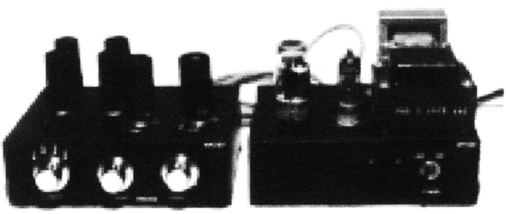
Where Push-Pull meets Single-Ended

WRIGHT Sound Company



The **WLA10 line amp** with 4 inputs for those who don't need the phono section. Dubbed by those that have listened, as the best sounding line amp they have ever heard, tube or solid state. All this for $365.00 plus $17.50 shipping and handling in the continental U.S., WA res. please add 8.2% sales tax.

The **WPP100A phono preamp**, the most natural sounding unit on the market today, is available at $529 plus $17.50 shipping and handling in the continental U.S., WA residents please add 8.2% sales tax. The WPP100A has gold RCA connectors, and a new WPS02 power supply with a power switch and plate and filament indicators. The performance is better than the original version, which beat all the competition in listening tests by members of **VALVE** and other audiophiles who have had the pleasure of reviewing this product.



Now available to **VALVE** members, and those who have tried S.E.X. amps, the **WPL10V complete line amp/ phono stage** component. This basic model has the quality of the WPP100A, with the additions of a selector switch with phono plus three other line inputs and volume controls to make this the center of that great new S.E.Xy sound system. No longer do you need to wait for a great sounding addition to have great S.E.X., and at just $649 U.S. funds plus $17.50 shipping and handling, you can get this fully assembled preamp/line amp delivered to your door in the continental U.S., WA residents add 8.2%. The WPL10V is designed to be a cost effective basic chassis type, constructed with all the great stuff that goes into the WPP100A. We made it especially for you S.E.X. owners and **VALVE** members who want the most out of your system for the least out of your pocket. I must add that this product will work with almost any power amp you now have or may purchase, so with or without S.E.X. this is a great addition to the WRIGHT line. Stay tuned for future models.

Please send your order and payment to:

WRIGHT Sound Company

3516 So. 262nd, Kent WA 98032-7047

For further information, please leave a message at (206) 859-3592

Please note: These items are individually hand built, and current high demand can

different circuits, to judge for myself.
First, naturally, I subbed the nickel TFA-204, a cap (more on this in a minute), and a 30H choke from the B+ line in for the standard TFA-204 in my 'slightly' tweaked S.E.X. amps.
Woah.
Something good is going on here.
Very good.
I had to call Eileen down to listen to see if I was hallucinating.
Now Eileen could really care less about how my stuff sounds. She says it all sound good, because she doesn't hear much difference.
But not this time.
"Wow, what did you do? Those voices sound more natural than they ever have."
By Eileen's standards, this rates in the Order of Magnitude Change category.
Okay, I'm not crazy. The midrange is smooth like Baby Katlyn's butt, and I've never heard so much bass detail and speed through the Whamo subs. Top end is definitely way smoother too. Even a crusted out LP with strings that feel like a bent back fingernail sounds vastly improved.
The midrange sweetness of the TFA-204 is there, but much more subtle, cleaner, a more refined balance.
I started with the coupling cap between the plate and the OT primary, then moved it to the ground leg of the OT to see if it made a difference. It made a definite improvement in detail and clarity. Thanks, Sandy.
I then tried the M6 steel lam version and got some of the benefit, but the incredible "jump" of the nickel was lacking. I couldn't go back.
Next we put this setup in our Afterglow prototype. Similar sonic improvements, although a bit bash shy. The circuit no doubt needs tweaking in any amp you splice it into.
But the same improvement in bass speed, and mid and treble clarity was there.
At this point I decided to see if I could find some measurable difference between standard direct coupled air gapped iron and parallel fed nickel.
At 1 watt output I got the following results Not insignificant!
Lesson 1
If you're going to take full advantage of parallel feed you should use nickel.

trans	frequncy resp.	THD@ 1kHz	3rd Harmonic
gapped iron	28Hz - 19.3 kHz	2.13%	32 dB down
interl. nickel	24.6 Hz - 26 kHz	0.7%	45 dB down

Oh yeah, about that coupling cap. I talked to Mike, and he feels the best way to calculate it is to get one of the reactance-frequency nomographs (try *Audio Cyclopedia,* p.1679) that has inductance on one 45 degree axis and capacitance on the other. You look up the inductance of your plate loading choke, and then the roll off frequency you desire on the horizontal axis. At their intersection locate the intersecting capacitance value. Mike suggests a frequency of about 15-20Hz. I used 10Hz to make the look up on the squinty chart more easy.
For my 30H choke I get about 7.5 mfd.
For Mike's new 50H, 60 mA choke I get about 4 mfd.
I learned that the high frequency extension needs some care as well as the low, and ended up paralleling a 1 mfd Siderial and a .1 mfd Teflon cap to a 10 mfd oil cap.
We also tried a 6.8 mfd Black Gate with good results, and an Axon and a Solen with mediocre results.
Lesson 2:
If your going to take full advantage of parallel feed, you have to use a very good coupling cap.

I started thinking about the 8 watt rating of the standard airgapped TFA-204, and asked Mike if I might get away with more power through the cap coupled nickel version.
"Yes", he says, "but not a lot more. Maybe ten watts. Okay maybe twelve, but I'm tellin' you Schmooley, you'll be pushing it."
See, there's no DC, but the with nickel you need a core cross section twice as big as iron for the same power, so in theory you still need a core area about the same size with a parallel fed nickel trans. Well, I needed to come up with a circuit to try this out. No 300Bs lying around, so I needed to figure out another way to use the 3K primary of the "nickel critter" to get 10 or so watts.
A little study of the tube manuals showed that a lot of wierd combos involving running big ass transmitting triodes at low B+ might work.
I tried an 845 running at 500V after JC Morrison's budget Dinosaur in *SP*, but the 845 I had was pooped out.
So I start looking at the 50 curves in the same page, and I bink on the fact that parallel 50s would love 3K.
Cripes Bottlehead, who's got 50s?
We'll me for one. Actually me for four used 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 cou-

pled 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.

Lesson 3:

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

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

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)

JF