Audio Classroom

Designing Your Own Amplifier, Part 3: Phase Inverters

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Most modern amplifiers have push-pull output stages, for reasons that will be more apparent when we discuss that part of the design. With push-pull output it is possible to get much more power from tubes and other components of reasonable size, with much less distortion. But before we can operate output tubes (or any other stage) in push-pull, we must convert the single-ended signal handled by most voltage amplifiers to the pair of oppositely phased signals needed for push-pull drive. This can be achieved with a *phase inverter* or *phase splitter*. Two principal kinds of circuits can be used: the transformer phase inverter, or the various types of tube inverter. Before going into their design, their relative merits should be stated briefly.

The simplest circuit to design uses transformer coupling. In the early days, transformers were not very good—they introduced a significant part of the overall amplifier distortion. Modern transformer materials and design have remedied this situation, so that transformer coupling can now be the best as well as the simplest, except for one serious obstacle: the problems in applying feedback over an amplifier using interstage transformer coupling are much greater than with RC-coupled types. For this reason, with the current trend toward amplifiers having large amounts of feedback, it is not surprising to find that interstage transformers are little used. Furthermore, good transformers are costly. But to obtain the same signal-voltage swing from RC-type tube inverters, a greater B+ supply voltage is invariably necessary, as will become evident later in this article.

SERIES TRANSFORMER

The merit of the transformer circuit is the simplicity with which two voltages of opposite phase can be provided on its secondary by the use of a center tap, as shown in *Fig. 1*. The reason why transformer coupling can handle a larger swing, using the same B supply, will become evident from *Fig. 2*, which shows characteristic curves of a 6SN7 tube (one half) or a 12H4, a miniature single triode having the same characteristics.

The dotted line in *Fig. 2* represents the DC voltage drop in the primary winding resistance of the transformer. This way, with a supply voltage of just over 260V, we can have 250V on the plate at 9mA plate current, and the voltage swing at the plate can be from 90V–400V—*much higher than the supply*. This is because, although the transformer has only a low primary *resistance*, it has a very high *impedance*. To obtain the necessary bias for the operating condition of *Fig. 2*, –8V, the bias resistor should be about 900Ω. The preferred value of 910Ω will be near enough.

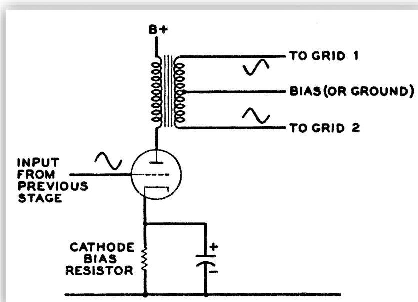


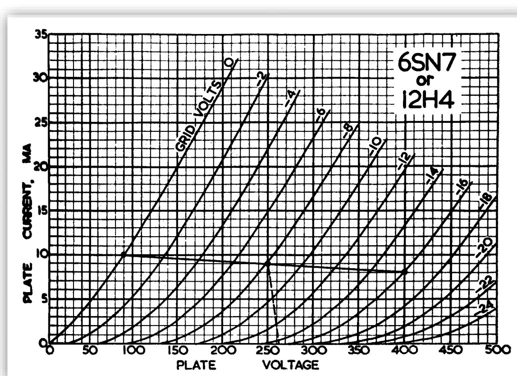
FIGURE 1: Circuit of phase inverter using a series or direct-coupled transformer.

The excursion shown in *Fig. 2* is 160V in the negative direction, and 150V in the positive. This is an off-center effect of 5V in a pk–pk swing of 310, representing only about 1.6% harmonic distortion.

Figure 3 shows one of the problems associated with this circuit. The primary inductance makes the load line open into an ellipse at lower frequencies. This not only reduces the maximum swing, but produces distortion quite rapidly. This ellipse—the widest possible before really serious distortion sets in—shows an AC voltage of 290 pk–pk for a current swing of 14mA, representing a reactance of 21k. To respond without distortion down to 20cps, the inductance would have to be 200H with 9mA flowing, which is quite an inductance! *[We have left the designation for frequency as it was in 1956, when this article was written. The cps, cycles per second, are now referred to as Hz (hertz). —Ed.]*

Reducing the current would be done by biasing the tube back a little further, and that would seem to make the inductance easier to get; but we lose more than we gain, because lowering the operating point on *Fig. 3* means that the width of ellipse permitted is narrowed, and we need much more inductance. The gain in inductance does not keep pace with need for more.

FIGURE 2: Load line representing one possible operating condition with series transformer coupling. The dotted line represents DC drop from the transformer primary-winding resistance; this operating condition requires a cathode-bias resistor of 910Ω, with a B+ supply that puts 250V on the plate.



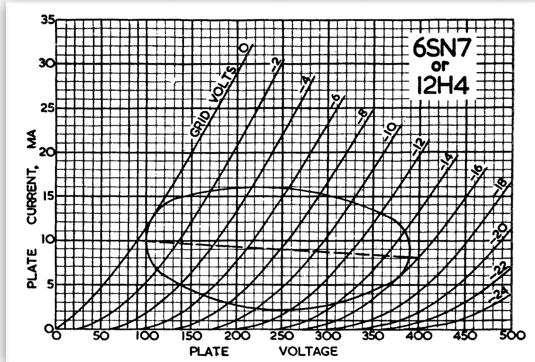


FIGURE 3: The ellipse approximating the load condition for minimum permissible shunt reactance has been sketched over the load line in *Fig. 2* (shown dotted).

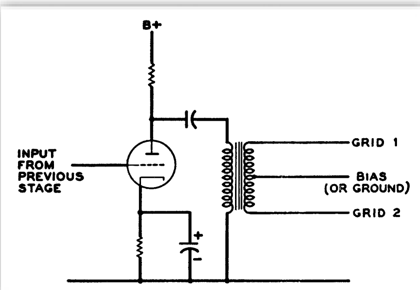


FIGURE 4: A parallel-fed transformer phase inverter. DC is kept out of the primary.

PARALLEL-FED TRANSFORMER

One way around this is to apply the parallel-fed method of connection shown in *Fig. 4*. That keeps the plate current out of the primary, so a transformer utilizing a better core material can be used. This will be a much more compact unit, with a considerably improved frequency response. Also, the possible improvement in design makes practical a transformer with a bigger voltage step-up from primary to secondary, so that the loss in signal-handling capacity resulting from the parallel-fed connection (because of reduced plate voltage) can be made up in the transformer.

But we can take the design of this stage as a starting point for considering the various RC-type tube inverters, which will make it easy to compare relative performance. *Figure 5* shows the load line for a plate resistor of 25k from a supply voltage of 275. Bias for maximum signal swing should be –7V, permitting a swing from 0 to –14V. Plate voltage at this operating point is 182, for a current of 3.8mA. This means the bias resistor should be 1,800Ω.

The pk–pk plate swing, from 75 to 265V, is 190V. This should be compared with 310V for the series transformer with a lower supply voltage. If the change permits us to double the transformer step-up ratio, we shall be better off. The distortion is not so good, however. Midpoint between 75 and 265 is 170V, so the swing is 12V off-center in 190, a distortion of about 6.3%.

The dotted ellipse represents a shunt reactance that permits a pk–pk current of 4.8mA at a pk–pk voltage of 180, about 37.5k. The inductance to maintain this to 20 cycles is 300H, which is reasonable with no DC in the primary.

SPLIT-LOAD INVERTERS

The 6SN7 is representative of tubes used for RC-type inverters, so we will continue making comparisons of circuits using this tube. All tube-type inverters effectively use one tube section to provide inversion without any gain. The difference lies in the method of sacrificing the tube gain to provide the reversal of phase, and the merits of each circuit are connected with this feature.

A transformer does have the advantage that if the turns on each half of the secondary are equal in the first place, nothing is likely to throw them out of balance subsequently. But using a tube as an inverter means that balance tends to be more or less dependent on the gain of the tubes. Some circuits avoid this deficiency, at the expense of something else.

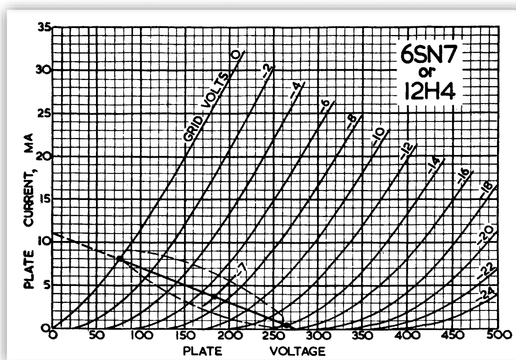


FIGURE 5: A typical load line for a parallel-fed transformer inverter, with a plate-load resistor of 25k and a bias resistor of 1,800Ω. B+ is 275V. Compared with series coupling, the maximum output swing is considerably restricted, but improved transformer design may compensate for this by providing increased step-up. The ellipse represents minimum permissible shunt reactance.

Probably the simplest type of RC phase inverter is made by simply putting half of the plate load in the cathode circuit, as shown in *Fig. 6*. It does not matter where the resistance in series with a tube is, so long as it has the same effect in restricting the plate-to-cathode voltage according to the current the tube draws. A load line for this method of operation can be drawn in exactly the same manner as for a normal voltage-amplifier stage, using a resistance value made up of the two resistances in series.

With the same operating conditions for the 12H4 (or half of a 6SN7) that we just considered for a parallel-fed transformer load, but putting 12.5k in the plate and 12.5k in the cathode, and with the same bias, we would have a total maximum swing of 190V pk–pk or 95V pk–pk for each output. In the more conventional RMS units, that is about 33.5V.

This is further restricted because we have to use grid resistors for the following tubes. Output tubes should have grid resistors of not more than about 250k. Using this value to construct a load line, the total shunt effect across 25k will be 500k, giving an AC load line of 23.7k. This is shown in *Fig. 7*. The maximum swing is between 79 and 260V, or 181V pk–pk, with a distortion of 6.9%. Although this is second-harmonic distortion, it is in the same phase on each output-tube grid and cannot be cancelled. Some reduction will occur because of feedback in the cathode of this tube, leaving about 1.59%, which is comparable with the transformer type, although the swing is much smaller.

There is a limitation with this kind of circuit: the output swing is not very great. For some output tubes requiring a bias less than 45V, this drive voltage would be adequate, but for other types, a phase in-

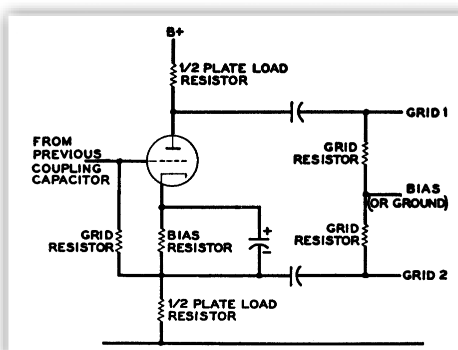


FIGURE 6: A split-load inverter. Significant components are marked on the diagram.

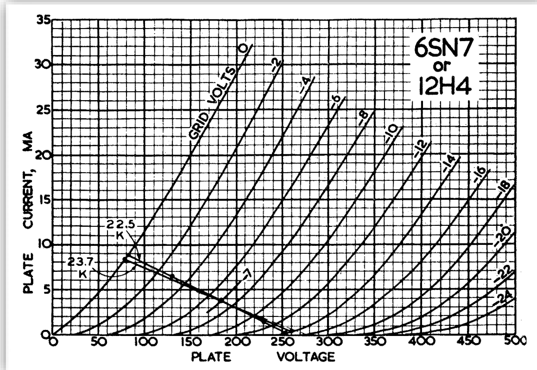


FIGURE 7: Load lines for the various RC phase inverters discussed in the text. The same operating point as given in *Fig. 5* is assumed, with 275V B+.

verter that gives a larger swing will be necessary.

This kind of circuit does, of course, avoid all the problems encountered in the transformer-coupled circuit. There are no shunt reactances of a value anything like as low as those we have been discussing. But another disadvantage of this circuit is that it has very little gain; in fact, as considered from the grid of the phase inverter to the grid of each output tube, it actually has a loss.

Assume that the circuit is being driven at full amplitude so we have 90V pk–pk being supplied to each output-tube grid. This 90V appears between cathode and ground of the inverter tube. To get this output in its plate circuit, the tube requires a *grid-to-cathode* voltage swing of 7V peak, or 14V pk–pk. This is over and above the voltage appearing at the out-

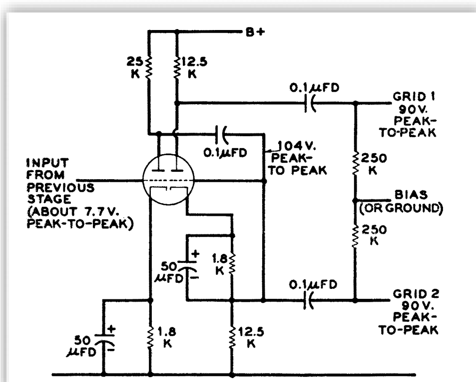


FIGURE 8: Complete double stage of the split-load inverter. The values shown are based on design figures in the text.

put from *cathode to ground*, so the total voltage necessary from *grid to ground* will have to be 104V pk–pk.

We need to put in 104V, then, to get out 90V for each output grid. It is quite easy to get a swing of 104V from a similar amplifying tube, operated single-ended; a suitable circuit is shown in *Fig. 8*. The first stage has a 25k plate resistor. A 470k grid return to the off-ground bias point is adequate for the second stage, or phase splitter.

But because this resistor is connected between points with a swing of 104 and 90V, respectively, it will draw much less current from the 104V circuit than if it were connected to ground. The voltage difference across it is only 14V, instead of 104V, so the current will be divided by 104/14. The *effective* AC resistance in parallel with the first stage's 25k plate resistor will be 470k multiplied by 104/14, which is about 3.5M. The shunting effect of this can be ignored.

Going back to *Fig. 5* to calculate the gain of this stage, a 14V swing gives an output of 190V swing—a gain of 13.5. So an output of 104V swing will require an input of 7.7V swing, or about 2.7V RMS.

The response of the phase splitter section will be 3dB down at the frequency for which the coupling capacitor reactance is equal to the grid resistors, 250k. With 0.1µF capacitors this is 7.5cps; good balance and response will certainly be maintained down to 20 cycles. For the previous stage, the *effective* grid resistor that determines response is 3.5M, which extends the response down to below 0.5cps with 0.1µF.

This kind of inverter gives good balance, because the balance is determined by the equality of the plate and cathode resistors. They can be selected precisely and can be expected to remain fairly well matched. Frequency response can be made as good as we want, but harmonic distortion is not as low as has been claimed, and the output voltage is limited.

If the kind of output stage we're thinking of needs more than 90V to swing each output tube, we have to consider another kind of phase splitter. Unless we are prepared to furnish an appreciably larger B+ voltage, the only way out is to use some kind of phase splitter in which each output comes for an entire plate load. This means we must use a separate plate circuit to drive each grid. If we can do this, using a 6SN7 tube and the same operating values we have already discussed, we shall get the full 190V (or nearly that) from *each* plate circuit.

PARAPHASE INVERTER

A simple system of using both plates for drive, known as a paraphase circuit, is shown in *Fig. 9*. We will continue using the same basic circuit values to get the advantage of relative performance comparison. But here the 250k output-grid resistors are coupled one to each plate, which brings the AC loads down to about 22.5k. This is shown in *Fig. 7* also. A maximum swing between 82V and 254V is obtained, using the same bias point and 14V grid swing. This is a gain of 12.3, and the distortion is about 8.1%, which is much higher than the split-load type ended up with.

In this case, however, each tube will produce second harmonics of opposite

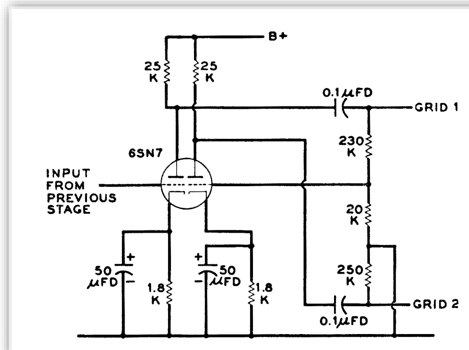


FIGURE 9: The paraphase circuit has a higher maximum output swing than the split-load circuit, because each output comes from an individual complete plate load.

phase, which will cancel at least partially in the output. At full load, the distortion in the output originating from the phase inverter would probably be about 3%. But the distortion will not be so high *at the same level* as in the previous circuit. The figures just discussed are for a swing of 172V. To get approximately the same output swing as the first circuit, we can use a grid swing from –3 to –11V (or, better still, from zero to –8V, using a new operating point). This gives 100V pk–pk output swing, and reduces the distortion on each side to 2%. The resultant in the output will be well under 1%.

To complete the design, though, we need a drive for the second tube section equal to that for the first. Since the gain is 12.3, we need to tap off 1/12.3 of the output from the first half, which will be in opposite phase to its input. The values shown give this, and provide the same total resistance in each grid circuit. Frequency response will be similar to that of the split-load circuit if similar values are used, so there is no need to discuss this again.

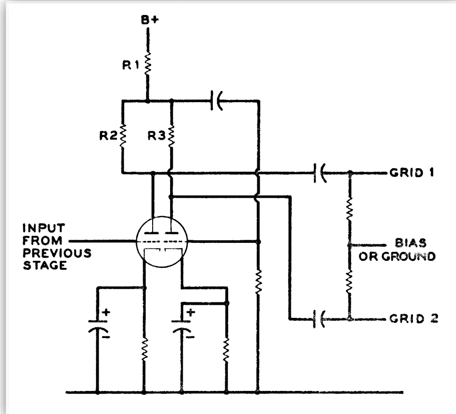


FIGURE 10: Modification of the paraphase circuit, intended to minimize any out-of-balance condition caused by variations in individual tube characteristics. This is called the floating paraphase circuit.

This kind of circuit will usually work quite well with a low-gain tube such as the one we have chosen. It is not too successful, however, when used with a high-gain tube in which the amplification factor is subject to considerably greater variation with operating conditions. This is particularly likely to happen if the heater voltage fluctuates, which alters the operating temperature of the tube and hence its space charge. Low-gain tubes are not as susceptible to this kind of variation.

The fact that the balance of the drive to the two halves of the output is dependent at all upon the exact gain of the tube is considered a disadvantage by some, for which reason methods have been devised to equalize the output so it is less dependent upon the gain of the individual tubes. We will not at this point go through the full design procedure, but merely indicate the method.

THE FLOATING PARAPHASE

Instead of using a predetermined tap in the grid circuit to obtain drive for the second stage, a common section of plate resistor is employed in the floating paraphase circuit in *Fig. 10*.

Plate currents of both tubes pass through R_1. The second tube needs a grid swing that is only a fraction of each plate-output swing to drive it, so as to give an output equal to that of the first tube. If both plates had exactly equal swing in plate *current*, there would be no swing across R_1; the swing at this junction point automatically adjusts itself so that the plate currents are out of balance by a fraction sufficient to provide the grid swing for the second tube. By adjusting the values of plate resistors R_2 and R_3, which are responsible for developing the actual output voltages, one stage can be arranged to have a gain larger than the other by this fraction.

Assuming that the gain of each stage can be 12, the voltage appearing at the junction must be 1⁄12 the voltage appearing at each plate. If R_1 is approximately equal to R_2 and R_3, then the plate-current swings will have to be in the ratio of 12:11. This means that to give equal plate-voltage *swings*, R_2 and R_3 must be in the ratio 11:12. Thus we can obtain equal outputs by altering the plate-load resistances only a fraction. This is a form of feedback, and we can expect the improvement to be about equivalent to the working gain of each tube.

Here is the rub: the series resistance R_1 drops some of the available B+ voltage, so that the load line starts from a lower point. Otherwise viewed, it could be said that only part of the active plate load is used. This has the effect of restricting the maximum available output in comparison with the straightforward paraphase circuit. Of course, the circuit must be tailored to the desired swing; if a swing of 25–30V RMS is adequate for each output tube, the arrangement can serve as is. Otherwise, higher B+ is needed.

This circuit provides distortion cancellation in the same way as the straight paraphase, and frequency response is calculated in the same way.

CATHODE-COUPLED INVERTER

Another circuit that works almost the same way is called the long-tailed or cathode-coupled phase inverter. Instead of putting the common resistance in the plate lead, we put it in the common-cathode return, as shown in *Fig. 11*. The signal voltage for driving the second half is developed across this common resistance; the direction of the voltage drive is reversed by having the second-stage grid effectively connected to ground through a large capacitor. Bias for both tubes is taken from a suitable point on the cathode resistor.

Apart from this, the circuit's operation and the available swing that it can give are quite similar to the floating paraphase. There is, however, the following difference: the voltage for the second grid is developed across R_1, which is a cathode load for the first half also. This being so, the input to the second grid is linearized by feedback, which does not occur with the floating paraphase. Accordingly the second-harmonic cancellation stands a chance of being a little better when this circuit is used as an inverter, although the gain is reduced.

OUTPUT IMPEDANCES

It is essential that the impedance presented to the output-tube grids by the drive stage not be very high. Voltage-amplifier tubes can have grid-circuit resistances on the order of a megohm, with an impedance or AC resistance of 100k or 200k. But output tubes must have much lower grid resistance than this, because of the bigger physical size of the tube and the fact that the grid current, although minute, is larger than in the voltage-amplifier tube.

As long as the grids of the output tubes are always operated in the negative region and large grid currents do not begin to flow, a grid resistance in the range of 100k to 250k, but never greater than 500k, is acceptable for stabilizing the bias. The driving impedance should be in the region of perhaps 20k, as typified by the inverter circuits we have discussed.

We shall find that in push-pull power-output stages, high power can be obtained much more easily if we "drive" the grids by making them go positive during part of the waveform. This means that we have to supply a current to the grid, as well as to maintain the potential swing required to get the right output waveform.

Otherwise stated, the stage driving the output must be capable of supplying power for grid current to avoid the kind of distortion that occurs when such power is not available. As soon as the

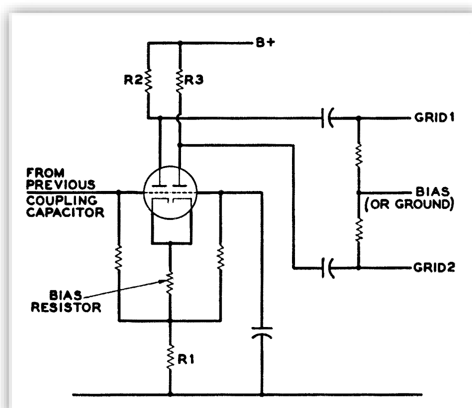


FIGURE 11: Similar to the floating paraphase is the cathode-coupled or long-tailed inverter. This is one variation.

drive reaches zero grid voltage, conduction of the grid in conjunction with the high source resistance prevents the voltage from swinging any further positive, and the waveform becomes "clipped."

Source resistances and impedances for drive stages have to be very much lower than when the same output tubes are operated without going into positive-grid region. But the design of a drive stage to supply power for the output grids is somewhat more complicated than the design of a simple power-output stage. We will reserve discussion of this until after we have covered the design of the push-pull power stage itself. Sometimes, also, to get sufficient swing for a large power-output stage, even without positive grid drive, a push-pull stage is used between the phase inverter and the output. This is called a drive stage, although not called on to handle power. ❖