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Application Note

SERVO MOTOR DRIVE AMPLIFIERS

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The design of transformerless, ac servo amplifiers using power darlington transistors and IC op amps are discussed. Two types of power amplifiers are illustrated, one using single +28 V power supply, the second using high voltage transistors in complementary configuration for operating directly off the line.

Four different op amp preamplifiers and 90° phase shifters are also described.

MOTOROLA Semiconductor Products Inc.

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INTRODUCTION

In many servo systems where an electrical signal has to be translated into mechanical motion, two-phase induction servo motors come into wide use. This mechanical motion of the motor is obtained by the rotating magnetic field derived from the relative quadrature excitation of the two field windings, the reference phase (fixed phase) and the control phase (variable phase). To produce this rotating field, and thus motor rotation, the two windings must be powered by signals 90° out of phase to each other. Either winding can have this phase shift.

One method of phase shifting the signals is to resonate the control phase with a parallel capacitor and shift the reference phase 90° with a series capacitor. Another method is to use two capacitors in series/parallel configuration to the reference phase to develop the relative 90° phase shift. The disadvantage of these techniques is that it requires relatively large capacitors (due to the low motor impedance) that must be able to sustain peak line working voltages.

A third method is to produce the 90° phase shift in the servo amplifier and energize the reference phase directly from the line, thus eliminating the large capacitors. This phase shift can be obtained at a low level and high impedance resulting in a small capacitor.

This is the method used in the preamplifiers designed for this application note. The servo amplifiers were designed for 60 Hz operation but can be scaled up to 400 Hz by changing the 90° shift capacitor.

Two basic servo power amplifier designs are illustrated in this report. The first uses a single, relatively low voltage power supply for push-pull driving a low voltage, center-tapped servo motor control phase. A commonly used power supply/servo motor combination is +28 Vdc and a 36 V center-tapped control phase motor. These are the values chosen for the first servo power amplifier design. However, any voltage rating center-tapped motor, with its appropriate power supply, can be used when the amplifier is properly scaled to these factors.

The second power amplifier features a line operated, high voltage, complementary output design for direct driving a 115 V rms servo motor. This circuit requires two supplies, ±150 Vdc.

SERVO SYSTEM

A block diagram of the low voltage servo amplifier is shown in Figure 1. It consists of a 90° phase shifter, a preamplifier with approximately 40 dB gain followed by a push-pull Class B power amplifier with approximately 23 dB gain. The servo motor control phase is driven directly from the transistor collectors of the power amplifier, thus eliminating the output transformer with a resultant savings in space, weight and cost. The block diagram illustrates the complete servo system feedback loop with the motor shaft mechanically coupled to the follow-up potentiometer in the error detector. This elementary system could also include more elaborate closed loop schemes; however, no attempt was made in this application note to close the loop and measure servo system performance. The servo amplifier and motor were operated in the open loop mode only. Amplifier gains and bandwidths were fixed but can readily be changed by simple resistor, capacitor changes to satisfy particular system criteria.

PUSH-PULL POWER AMPLIFIER, LOW VOLTAGE SERVO

The power amplifier (Figure 2) is of a push-pull Class B, common emitter configuration using power darlington transistors for high current gains and input impedances compatible with that of the preamplifier output. The input is driven push-pull and is capacitively coupled to eliminate the input transformer; thus, the servo amplifier is completely transformerless. The darlington collectors are connected directly to the control phase, which when

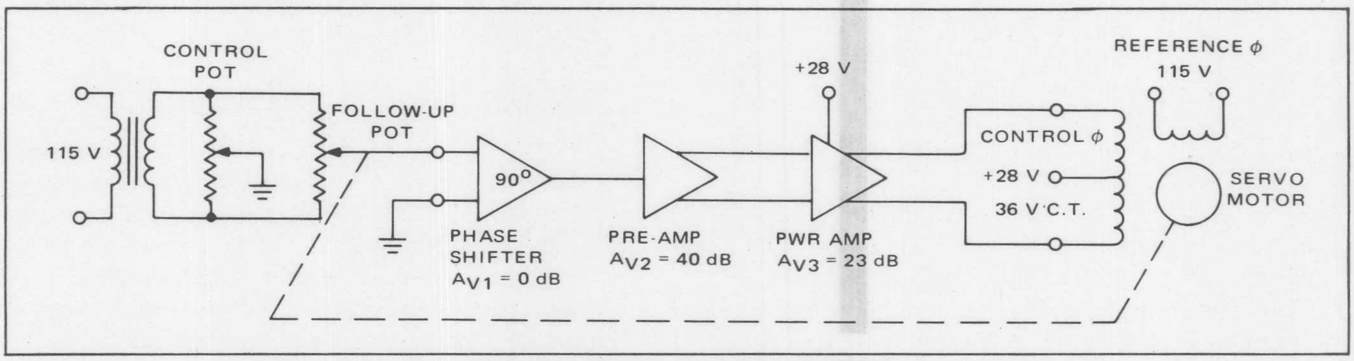
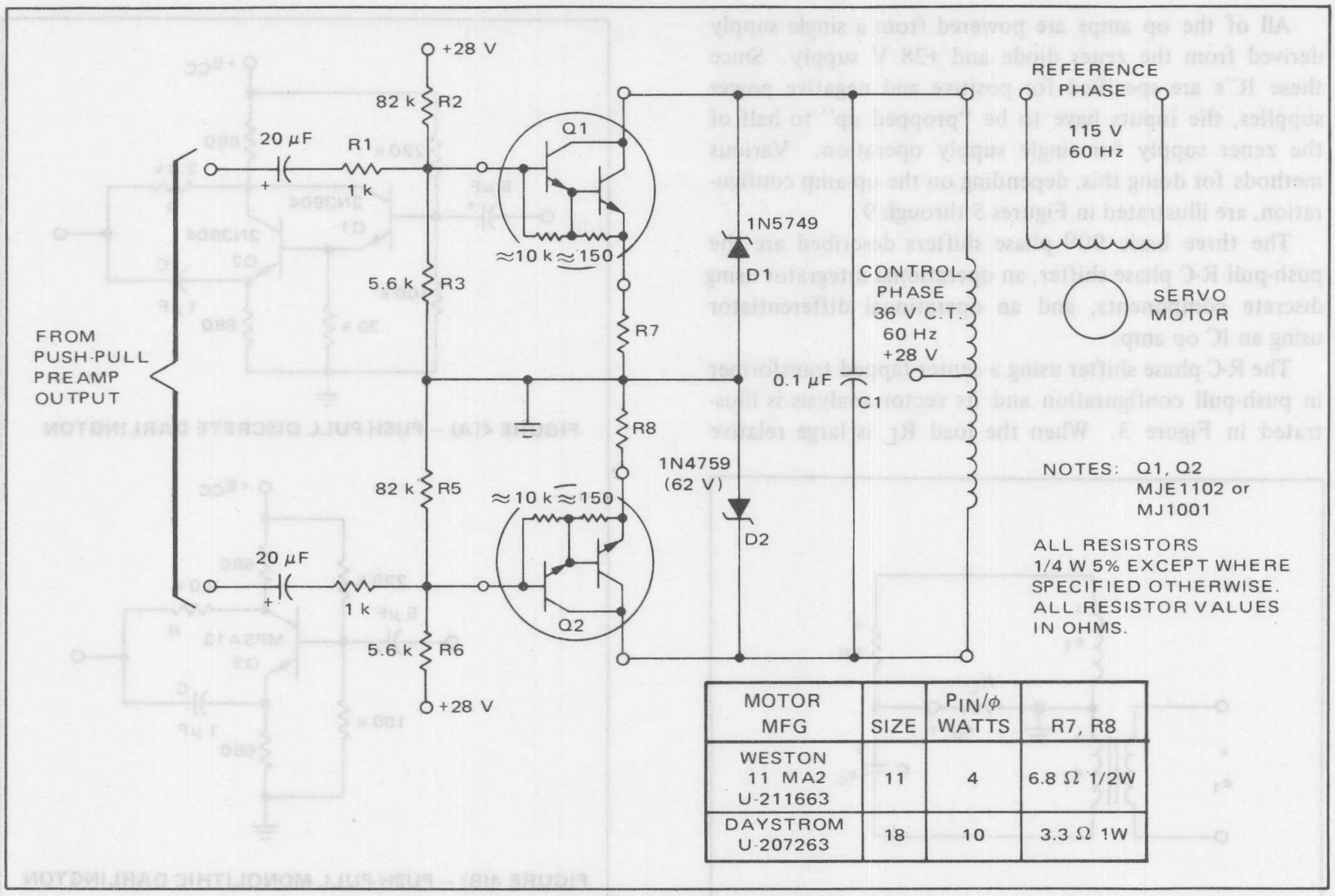


FIGURE 1 – SERVO AMPLIFIER BLOCK DIAGRAM



MOTOR MFG	SIZE	P_{IN}/ϕ WATTS	R7, R8
WESTON 11 MA2 U-211663	11	4	6.8 Ω 1/2W
DAYSTROM U-207263	18	10	3.3 Ω 1W

FIGURE 2 – SERVO MOTOR POWER AMPLIFIER

used with a 28 V supply, results in a collector/collector voltage swing of approximately four times the supply voltage less the transistor saturation voltages and emitter resistor voltage drops. This four times factor is the result of the auto transformer action of the center-tapped control phase (identical to the action of a center-tapped output transformer.) The control phase voltage therefore is approximately $\frac{4(28-2)}{2.828} = 36$ V rms.

The amplifier is actually biased slightly Class AB by the input bias network R2, R3, R5 and R6 to minimize crossover distortion. Quiescent, no-signal current will vary from approximately 10 mA to 20 mA depending on the emitter resistor and motor size chosen. Zener diodes D1 and D2 prevent over-voltage damage to the transistors by suppressing the inductive voltage spikes created when the amplifier is over-driven. Capacitor C1 limits the amplifier's bandwidth and prevents high frequency oscillation.

The power amplifier is designed to drive two different servo motors, sizes 11 and 18 (BuOrd) requiring control-phase power inputs of approximately 4 and 10 watts respectively. The only difference in power amplifier design for the two power levels is the size of the emitter resistors. Two different but similarly rated power darlington transistors were used for both motors, a plastic packaged MJE-1102 and a metal packaged MJ1001. These transistors must be capable of sustaining voltages greater than two times the supply voltage. These devices have BV_{CEO} of 80 V and minimum h_{FE} of 750 and 1000 respectively at collector currents of 3 A. Peak currents under load and hard over-drive conditions were approximately 250 mA and 650 mA repsectively with total circuit average currents of approximately 200 mA and 500 mA respectively.

Heat sinking of the darlington transistors is required for continuous operation to insure safe transistor junction temperatures.

PREAMPLIFIERS AND 90° PHASE SHIFTERS

Four preamplifier designs are illustrated using various integrated circuit operational amplifier configuration. Several different 90° phase shift circuits and their variations are also shown in Figure 4. These circuits can be used in any phase shifter/preamplifier combination.

Since the power amplifier is driven differentially, the preamplifier outputs must accommodate this requirement. When using a single ended op amp output, the signal must be phase-split as shown in Figure 7. This is accomplished by the simple phase-splitting amplifier Q3 where the equal amplitude outputs are taken respectively from the emitter and collector, 180° out of phase. This circuit has unity voltage gain (0 dB).

All of the op amps are powered from a single supply derived from the zener diode and +28 V supply. Since these IC's are specified for positive and negative power supplies, the inputs have to be "propped up" to half of the zener supply for single supply operation. Various methods for doing this, depending on the op-amp configuration, are illustrated in Figures 5 through 9.

The three basic 90^{o} phase shifters described are the push-pull R-C phase shifter, an operational integrator using discrete components, and an operational differentiator using an IC op amp.

The R-C phase shifter using a center-tapped transformer in push-pull configuration and its vector analysis is illustrated in Figure 3. When the load R_L is large relative

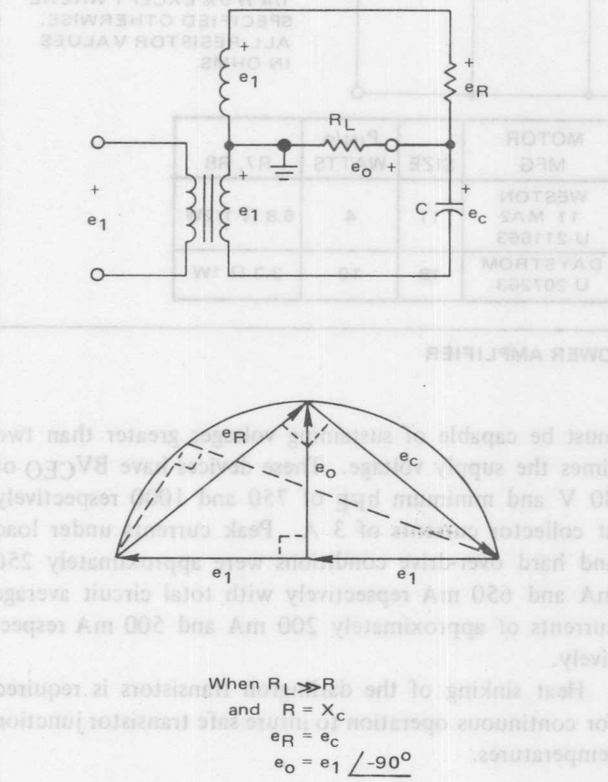


FIGURE 3 – PUSH-PULL RC PHASE SHIFTER

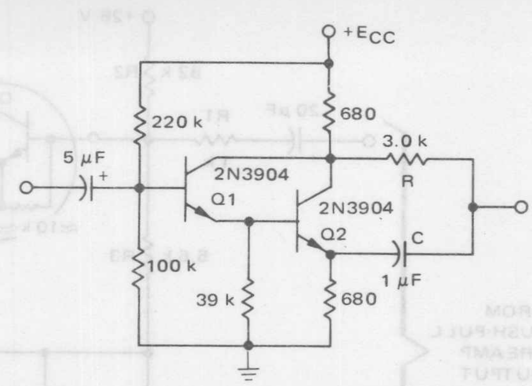


FIGURE 4(A) – PUSH-PULL DISCRETE DARLINGTON

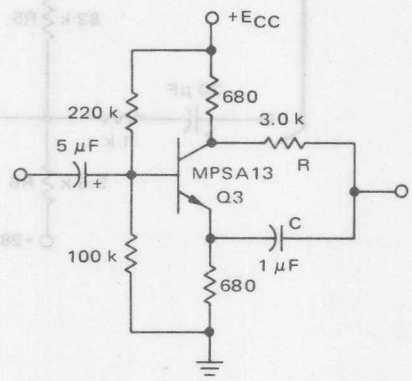


FIGURE 4(B) – PUSH-PULL MONOLITHIC DARLINGTON

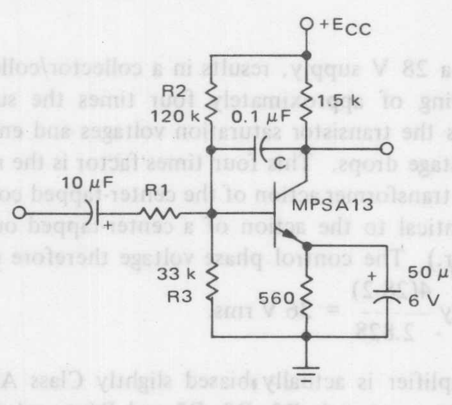


FIGURE 4(C) – OPERATIONAL INTEGRATOR

FIGURE 4 – 90^{o} PHASE SHIFTERS

to the impedances of R and X_C, the locus of the output e_o will describe a semi-circle whose radius (e_o) is a constant amplitude, equal to the input e_1, and whose phase angle is a function of R and X_C. Thus, by varying R, a phase shift from approximately 0^{o} to 180^{o} can be obtained at a constant amplitude. By making $R = X_C$, a 90^{o} phase shift will be obtained with 0 dB voltage gain.

Figures 4A and B show the actual circuit of this phase shifter using a phase-splitting amplifier for generating the push-pull outputs. It utilizes the darlington transistor configuration of two 2N3904's (Q1 and Q2) or an MPSA13 (Q3) which is a monolithic darlington. Due to the loading of the following op amp, the value of R was found to be somewhat greater than $X_C = 2.65\ k\Omega$ when $C = 1\ \mu F$ at 60 Hz. For 400 Hz operation, C should be proportionally reduced.

An example of a simple operational integrator circuit using discrete components is shown in Figure 4C. To obtain the large circuit gain required, an MPSA13 darlington transistor is shown, however, the two cascaded 2N3904's as previously mentioned can also be used. An operational integrator utilizes capacitive feedback to produce the following voltage transfer function:

$$A = \frac{e_o}{e_{in}} = \frac{-jX_c}{R} = \frac{1}{j\omega RC}$$

The output voltage will be related to the input by

$$\frac{1}{j\omega RC}$$

where the j operator produces the 90^o phase shift. When $R = X_c$, the voltage gain will be unity (0 dB), and will remain such as long as the frequency remains constant. For 60 Hz operation, with $R = 27\ k\Omega$, $C = 0.1\ \mu F$ (X_c = $26.5\ k\Omega$). To scale the circuit for 400 Hz, and still maintain unity gain, C should be reduced accordingly.

The third 90^o phase shifter, the operational differentiator, will be described in the following section illustrating preamplifier #4.

Preamplifier #1 (Figure 5) uses the 90^o operational integrator driving a MC1420 differential input, differential output op amp connected in an inverting configuration. With a single ended input, a push-pull output results which drives the darlington power amplifiers. Voltage gain is set by resistors $R1 = 10\ k\Omega$ and $R2 = 1\ M\Omega$ resulting in a gain of approximately 38 dB (40 dB theoretically with infinite loop gain). The op amp has excellent operating point stability due to the high dc feedback. The bandwidth is kept relatively narrow (approximately 4 kHz) for servo system stability by the 510 pF compensation capacitors. This op amp is specified for dual supplies of ±8 V maximum. However, for this single supply configuration, the supply is regulated to +12 V by zener diode D1. Single supply operation is accomplished by the resistor divider R3 and R4 feeding the differential input through R5 and R6.

Another method of obtaining the required differential output is illustrated in Figure 6, preamplifier #2. Here,

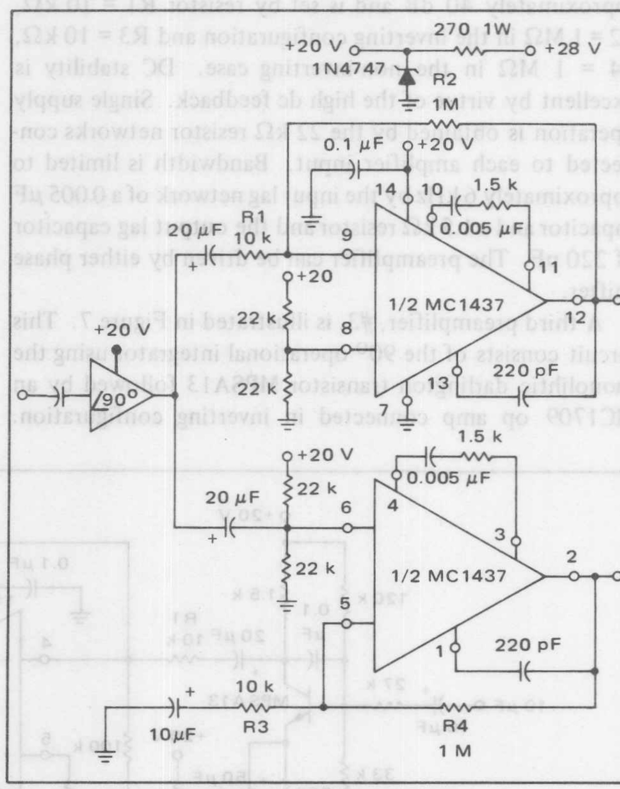


FIGURE 6 – PREAMPLIFIER #2

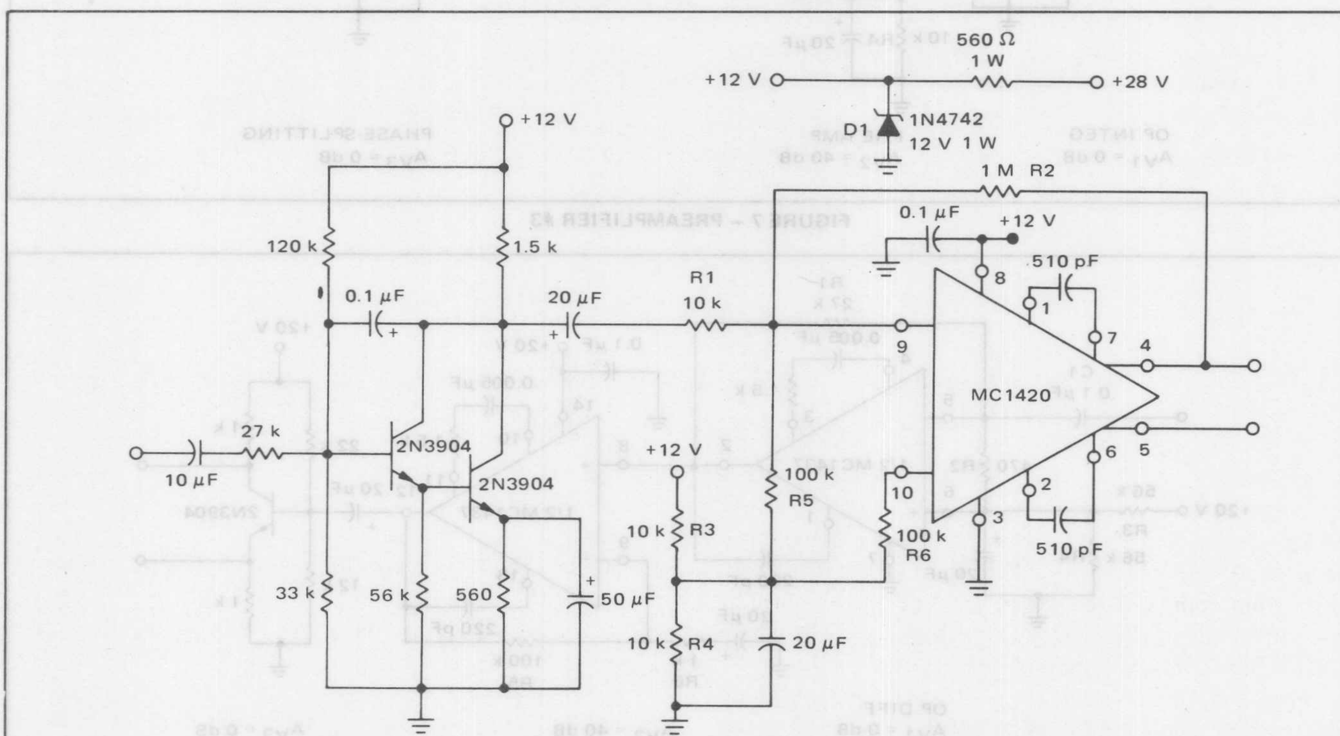


FIGURE 5 – PREAMPLIFIER #1

two op amps are connected in parallel from a common input, one in inverting configuration and the other, non-inverting. The respective outputs will therefore be complementary. A dual op amp, MC1437, is powered by a single +20 V zener regulated supply. This device can sustain a dual supply of ±18 V max. The voltage gain is approximately 40 dB and is set by resistor R1 = 10 kΩ, R2 = 1 MΩ in the inverting configuration and R3 = 10 kΩ, R4 = 1 MΩ in the non-inverting case. DC stability is excellent by virtue of the high dc feedback. Single supply operation is obtained by the 22 kΩ resistor networks connected to each amplifier input. Bandwidth is limited to approximately 6 kHz by the input lag network of a 0.005 μF capacitor and a 1.5 kΩ resistor and the output lag capacitor of 220 pF. The preamplifier can be driven by either phase shifter.

A third preamplifier, #3, is illustrated in Figure 7. This circuit consists of the 90° operational integrator using the monolihtic darlington transistor MPSA13 followed by an MC1709 op amp connected in inverting configuration. The single-ended output is complemented by the phase-splitting amplifier to push pull drive the power amplifier.

Voltage gain is approximately 39 dB as set by resistor R1 and R2. Again, single supply operation is accommodated by resistor divider R3 and R4. Bandwidth is approximately 4 kHz as set by the two lag networks of 0.005 μF and 1.5 kΩ and 100 pF respectively.

Preamplifier #4 is shown in Figure 8. It consists of an IC op amp connected as an operational differentiator followed by a direct coupled, noninverting op amp. The two op amps are contained in the dual op amp, MC1437. The differentiating network consists of the feedback resistor R1 = 27 kΩ and the differentiating capacitor C1 = 0.1 μF. This circuit was chosen rather than the integrator version since this configuration offers excellent dc stability by virtue of the high dc feedback of the 27 kΩ feedback resistor working into the amplifier's common mode input impedance. Single supply operation is accommodated by the resistor divider R3 and R4 which sets the input bias to half supply. The output of the differentiator is

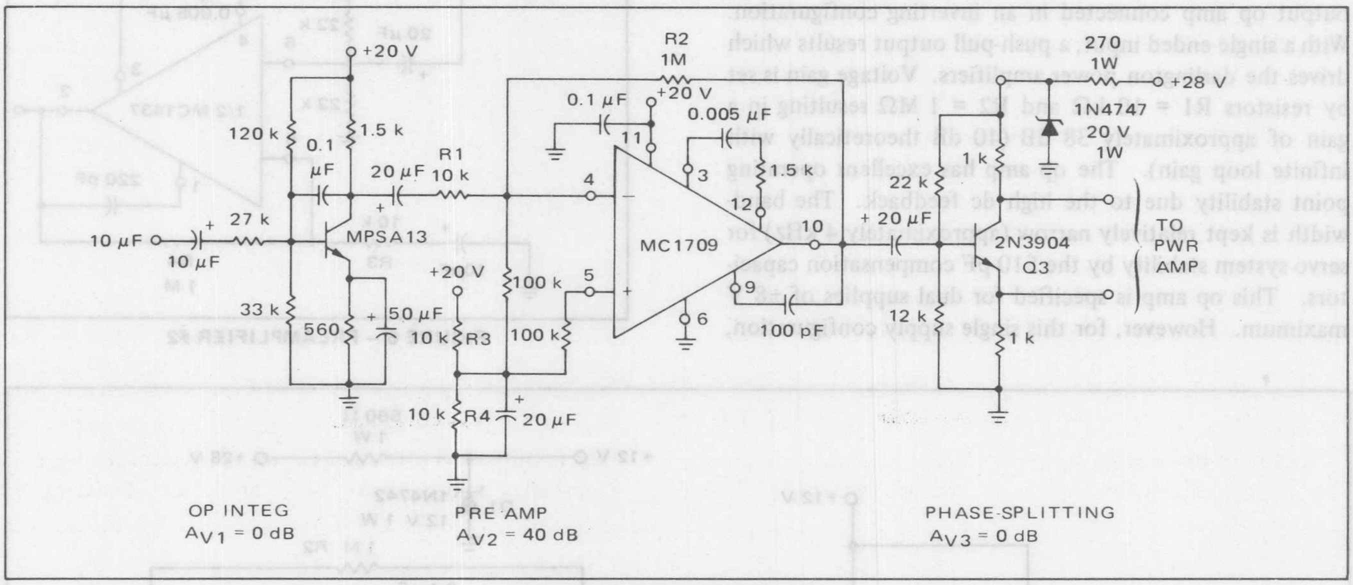


FIGURE 7 – PREAMPLIFIER #3

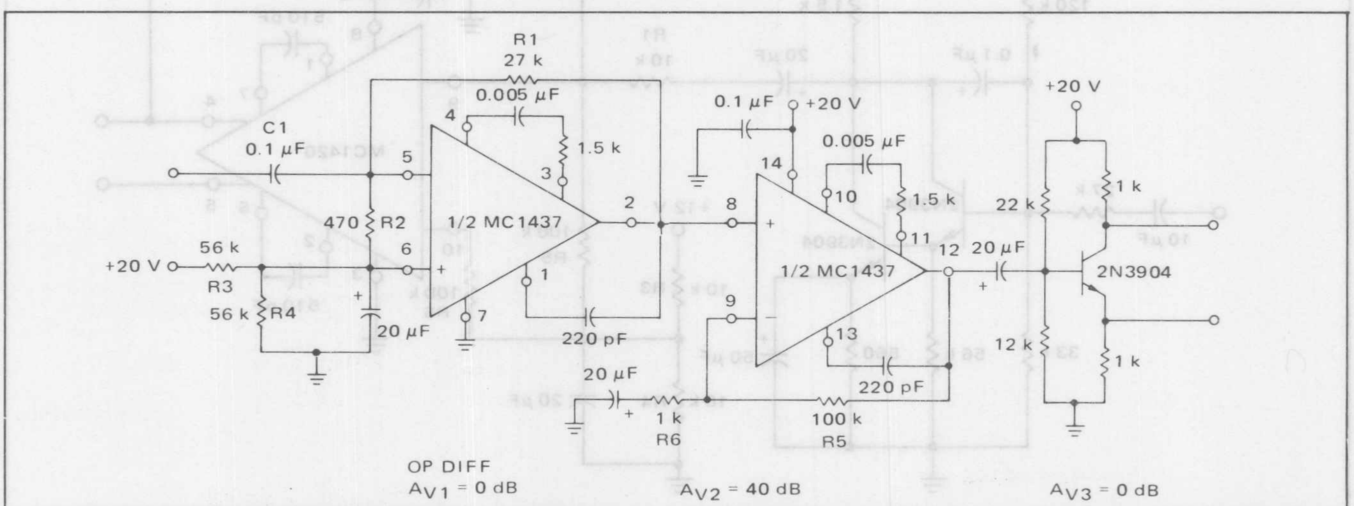


FIGURE 8 – PREAMPLIFIER #4

also at half supply due to high dc feedback and since it is direct coupled to the following op amp, it furnishes input bias to that amplifier. This op amp's ac gain is set by R5 = 100 kΩ and R6 = 1 kΩ, and has excellent output stability set at half-supply by the high dc feedback. The single-ended amplifier output is converted to push-pull by the following phase-splitting amplifier.

The voltage transfer function of the differentiator is $A = j\omega RC$ resulting in an output leading the input by 90°. When $R = X_C = \frac{1}{\omega c}$ the gain is unity. For 60 Hz, this occurs when C = 0.1 μF. To operate at 400 Hz, C should be scaled down to approximately 0.015 μF. Since the gain of the differentiator increases with increasing frequency, the circuit would tend to oscillate if it were not bandwidth limited. The indicated lag networks and the input shunt resistor R2 = 470 Ω stabilizes the circuit against oscillations.

All of the preamplifiers and phase shifters were temperature tested between 0°C and 55°C and preamplifier #3 tested between –40°C and +85°C. The maximum ac gain variation for any of the preamplifiers were less than ±3%. Additionally, all circuits exhibited excellent dc stability.

LINE OPERATED SERVO AMPLIFIER

There are some applications where it may be desireous to operate the servo system directly off the line using 115 V rms servo motors. A dc power supply is still required for powering the amplifier but it need only be of a simple design. Figures 9 and 10 illustrate just such a servo amplifier and power supply for direct line operation.

The amplifier consists of three blocks: the push-pull R-C phase shifter, the single-ended op amp preamplifier and the push-pull, Class B power amplifier. The first two blocks have been previously described. The power amplifier is of a complementary transistor design using high voltage (300 V) transistors, in a common emitter, darlington configuration. This circuit has adequate power gain to be driven from the preamplifier output.

When using balanced, plus-and-minus power supplies, the amplifier output can be directly coupled to the ground-referenced, servo-motor control phase. Thus, the coupling capacitor can be eliminated since the amplifier quiescent voltage is zero volts.

To develop the 115 V rms control voltage (162 V peak), dc power supplies of at least ±162 V are required. These supplies can be simply derived by rectifying the line voltage, neglecting rectifier drops and supply regulation, as illustrated in Figure 10.

Of the two supplies illustrated, the one using the power transformer (Figure 10A), is recommended, particularly in regards to the efficiency of the +20 V supply that powers the low level circuits. If cost or size is a criterion, then the direct-line operated supply (Figure 10B) can be used. However, considerable power is consumed in the +20 V series dropping resistors approximately (160 V - 20 V) (25 mA) = 3.5 W.

The darlington transistors, in the off state, must be able to sustain twice the supply voltage. The voltage breakdown ratings, BV_{CEO}, for transistors Q2 and Q4 are 300 V, and for Q3 and Q5, 350 V. This thus limits the permissible voltage levels of the supplies to approximately ±150 V. When fully loaded with a size 22 servo motor, the supplies, due to their relatively poor regulation, will drop to approximately 136 V resulting in a greater safety margin on the transistors breakdown limits. However, this lower supply voltage results in a lower than rated ac signal to the motor – approximately 264 Vp-p (93 V rms). For most servo applications, this reduced voltage will suffice.

The push-pull, Class B power amplifier, due to its complementary configuration, requires a single ended input that is supplied by C1 and C2 to the two respective complementary inputs of the power amplifier. These capaci-

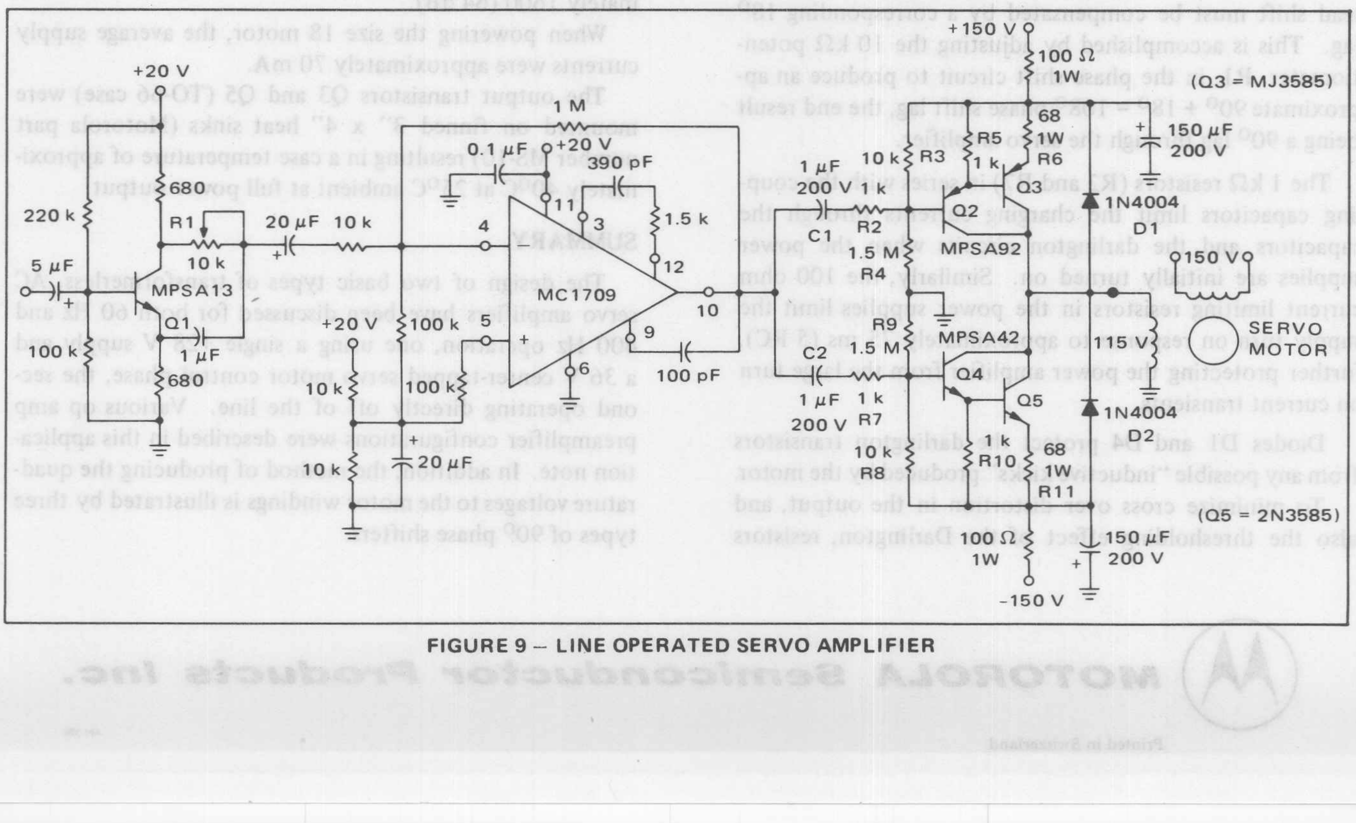


FIGURE 9 – LINE OPERATED SERVO AMPLIFIER

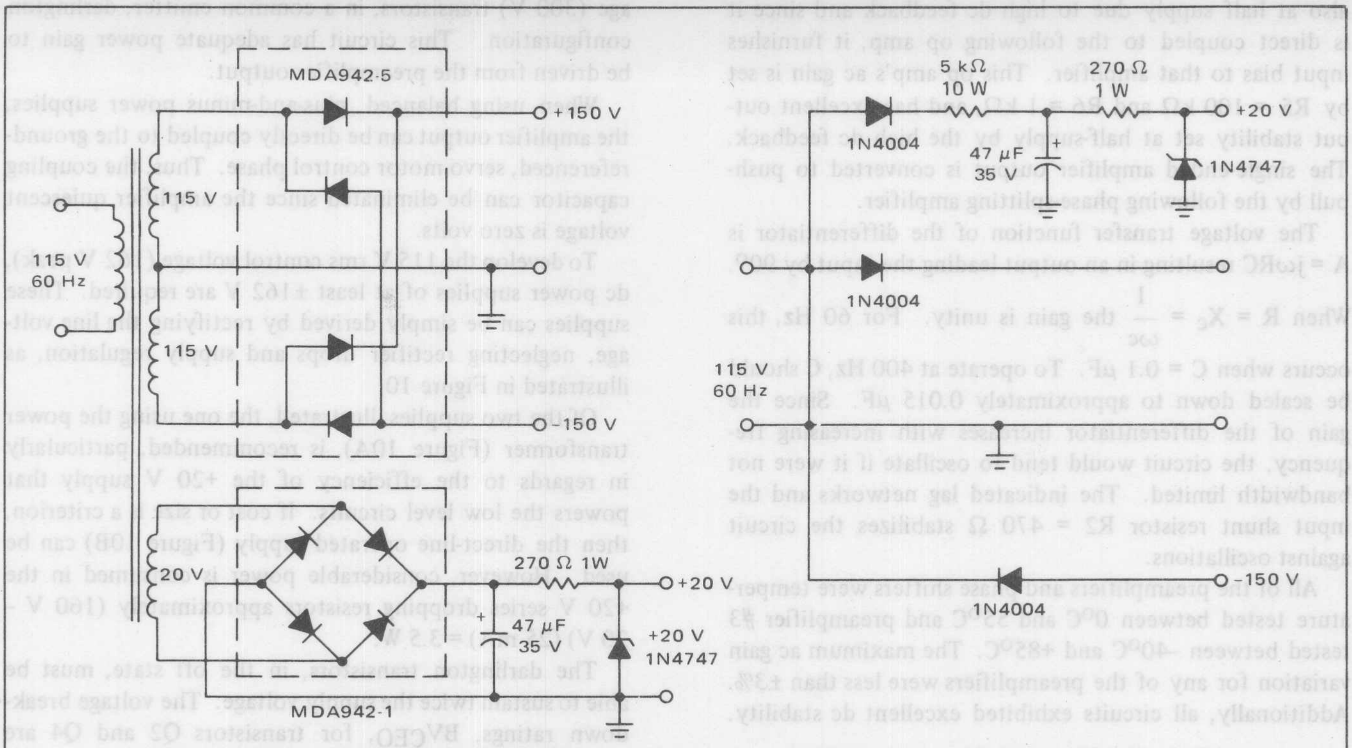


FIGURE 10(A) – POWER TRANSFORMER

FIGURE 10(B) – DIRECT LINE OPERATED

FIGURE 10 – POWER SUPPLIES FOR LINE OPERATED SERVO AMPLIFIER

tors must sustain voltages greater than the supply voltages and are therefore rated at 200 V. Due to size limitations, the capacitors were chosen to be 1 μF. The input impedance of each half of the power amplifier is approximately 8 kΩ which results in a phase shift of approximately 18° lead at 60 Hz through the coupling network.

Since the total phase shift through the servo amplifier must be 90° to maintain the quadrature relationship between the motor's control and reference phases, this 18° lead shift must be compensated by a corresponding 18° lag. This is accomplished by adjusting the 10 kΩ potentiometer, R1, in the phase shift circuit to produce an approximate 90° + 18° = 108° phase shift lag, the end result being a 90° lag through the servo amplifier.

The 1 kΩ resistors (R2 and R7) in series with the coupling capacitors limit the charging currents through the capacitors and the darlington circuits when the power supplies are initially turned on. Similarly, the 100 ohm current limiting resistors in the power supplies limit the supply turn on response to approximately 75 ms (5 RC), further protecting the power amplifier from the large turn on current transients.

Diodes D1 and D4 protect the darlington transistors from any possible "inductive kicks" produced by the motor.

To minimize cross over distortion in the output, and also the thresholding effect of the Darlington, resistors R4 and R9 are used to slightly forward bias the input transistors to approximately 0.7 V.

The voltage gain of the power amplifier using 68 Ω emitter resistors R6 and R11 and a size 18 motor is approximately 21 (26 dB). These emitter resistors, when beta multiplied by the darlington transistors, produce the power amplifier's input impedance of approximately 8 kΩ which is well within the drive capability of the MC1709 op amp.

The overall voltage gain of the servo amplifier is approximately 1600 (64 dB).

When powering the size 18 motor, the average supply currents were approximately 70 mA.

The output transistors Q3 and Q5 (TO-66 case) were mounted on finned 3" x 4" heat sinks (Motorola part number MS-10) resulting in a case temperature of approximately 40°C at 25°C ambient at full power output.

SUMMARY

The design of two basic types of transformerless, AC servo amplifiers have been discussed for both 60 Hz and 400 Hz operation, one using a single +28 V supply and a 36 V center-tapped servo motor control phase, the second operating directly off of the line. Various op amp preamplifier configurations were described in this application note. In addition, the method of producing the quadrature voltages to the motor windings is illustrated by three types of 90° phase shifters.