This article briefly collates design information relating to a vibrato effect used in Magnatone amplifiers in the late 1950's and early 1960's, and describes the design of the 213 model amplifier that uses the vibrato effect.

The performance of this vibrato effect circuit is specific to the model of Silicon Carbide voltage dependant resistor (varistor) used. Cloning of that varistor is of interest due to the cost of NOS parts.

Schober organs used this vibrato technique, starting with a variable triode resistance, then using varistors, and finally changing to LDRs. Although varistors weren't used, the technique was also used by others.

The Wurlitzer electronic vibrato technique was dominant at the time and is briefly noted.

Contents

Resistance-Reactance Bridge Vibrato Technique

In July 1953, Robert C. Moses [1] described [1] an oscilloscope phase-angle measurement technique that used a resistance-capacitance bridge to shift the phase of a signal passing through the circuit by varying R, as shown in the Figure 1 circuit.

Don L. Bonham filed patent # 3,146,292 in March 1954 where various phase shift circuit techniques, included Moses' circuit, cycle the phase shift through nearly 180 degrees at a low frequency, such that the signal frequency being passed is modulated by the low frequency to apply a vibrato effect to the signal.

Table 1 in Moses' article identifies the phase angle shift range achieved with the cathodyne based RC phase shifting circuit. Moses tabulated RC products, whereas Bonham's patent # 3,146,292 shows a simpler to comprehend plot of phase change for varying $R/X = R/(1/\omega C) = \omega RC$. This vibrato generation technique had the practical benefit that when the centre frequency (R/X=1) was set at about 1kHz, where the vibrato effect is maximum [2], then much lower and higher frequencies exhibit less vibrato (where vibrato is usually not wanted).

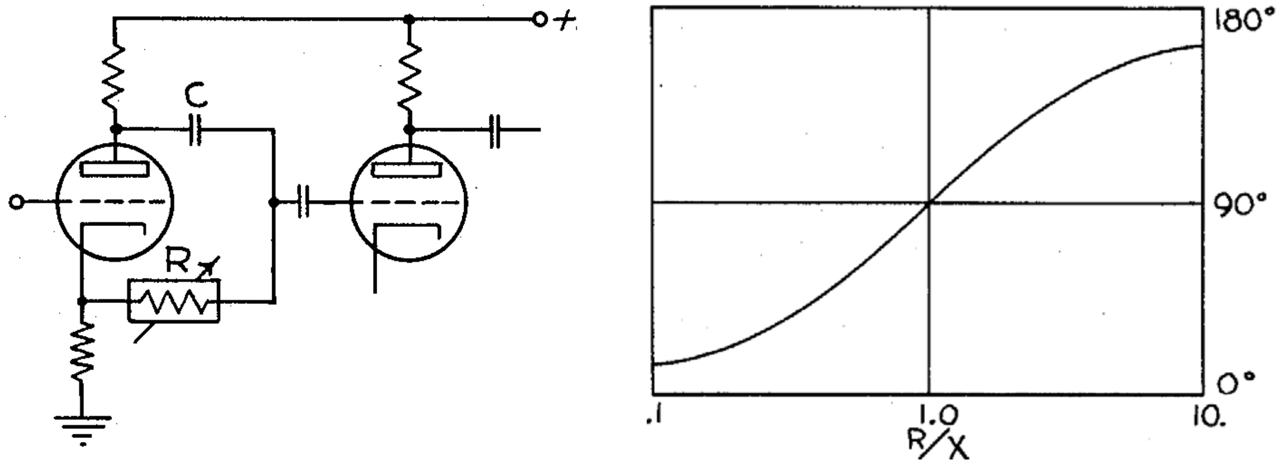

Figure 1. Moses' phase shift circuit, as Dorf and Bonham went on to use for vibrato generation.

Richard Dorf used this phase-shift technique in Schober organ kits from Feb 1956 [2], initially using a 6SL7 triode as the variable resistance arm of the bridge and a CT transformer winding to generate the out of phase

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[1] Moses appears to have worked for Sylvania in the late 1940s to early 50's, and then Lear Inc.

[2] A 1kHz signal will have its phase varied above and below 90deg by a greater level than say a 10kHz signal (R/X = 10) where phase will vary over a much small range around circa 160deg.

signals. The Schober circuit is somewhat complicated, perhaps as it had to circumvent Bonham's patent. Dorf's patent # 2,835,814 was applied for in March 1956, and covers that format, as does Dorf's Radio Electronics March 1957 article on phase swing vibrato, and the 1958 2nd Edition of [3]. Circa 1960, Schober's Concert and Consolette models, and the new smaller Spinet model, used varistors in circuitry similar to Magnatone's [see Appendix A. Schober vibrato circuits using varistors]. From 1963, Schober changed to LDRs in the variable resistance arm, and used transistor circuitry and incandescent bulbs to drive the LDRs [3]. The incandescent bulb and LDR devices have thermal and time delay characteristics that modify the phase modulation response, with a square waveform simply applied across the bulb.

Magnatone introduced this vibrato technique in 1957 to the Custom 200 series of guitar amps using a varistor circuit technique which Bonham filed patent # 2,988,706 for in Oct 1958. Some models used two sequential stages of phase shift, which allows a greater intensity of vibrato with less non-linearity introduced, as the phase shift in each stage isn't required to move as far to the 0 and 180 degree ends of the LFO sweep.

Introduced in 1961, the Hammond L100 organ (AO-41 module) uses a resistance-reactance bridge to generate vibrato, where a saturable inductor provides a variable reactance, with the resistance arm remaining fixed. The L100 uses three sequential stages of phase shifting and allows wet + dry mixing to give a chorus response. The X66 organ from 1967 also used saturable inductors in a 3 stage phase-shift vibrato for bass.

Ampeg used an LDR for the variable resistance arm in two sequential vibrato phase-shift stages in the Gemini GV-22 guitar amplifier introduced in 1968. Forrest Cook similarly used an LDR, and implemented digital LED waveform generation [9]. The basic phase modulator and LFO modulator scheme can also be replicated in solid-state circuitry by judicious design. Rod Elliot used n-channel FETs as the variable resistance device, and used an op-amp input to invert the phase for vibrato. Valves are replaced by JFET devices to allow a battery operated 'pedal', and the V-I curve of a diode operating at low current can be used to provide the large resistance swing needed for Moses' phase modulator in a pedal. Many variations using solid-state hardware implementation have been pursued for phase shifters and flangers and chorus effects for guitars [10].

The circuit diagram in Figure 2 from patent # 2,988,706 shows how Bonham used another cathodyne circuit to modulate the varistor resistance by varying the voltage across the series connection of varistors. The split signal input phases pass through capacitor C, and through the coupling capacitors and then through the varistors – with the combined signal output being capacitor coupled to the next stage. The varistors act in parallel to represent the resistance R in Moses' phase shift circuit.

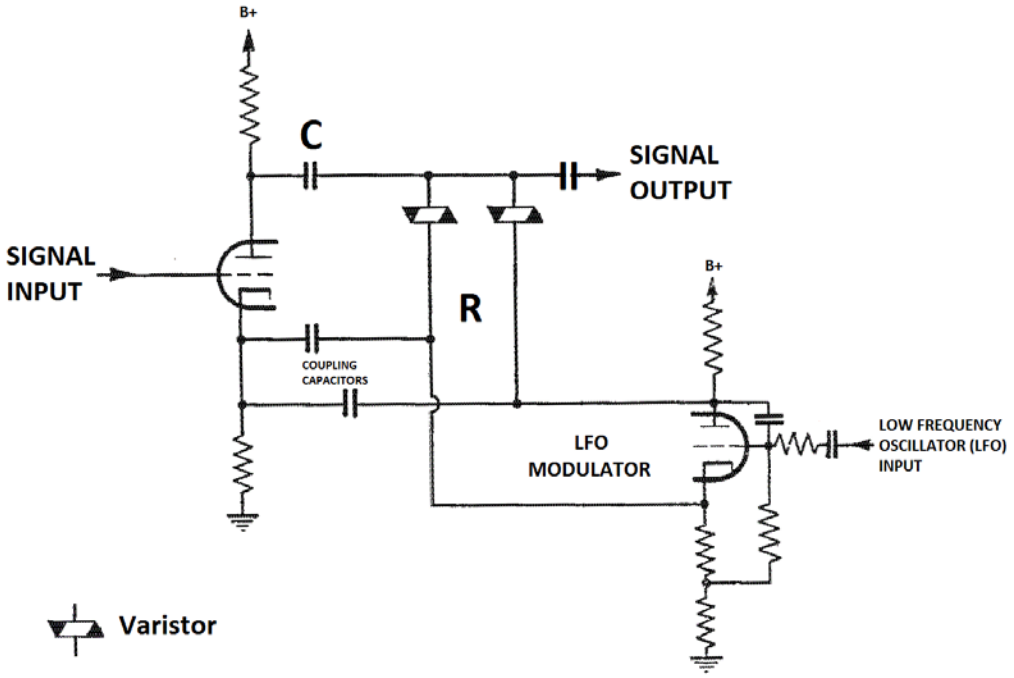


Figure 2. Magnatone vibrato circuit by Bonham

The Low Frequency Oscillator (LFO) signal drives the LFO modulator cathodyne circuit. The anode-to-cathode voltage of the modulator triode increases and decreases at the LFO frequency, which increases and decreases the voltage across the series connection of the varistors, which causes R to vary. The capacitor from anode to grid in the modulator stops any signal input from being amplified by the triode.

Magnatone and Bonham

By far the best reference for Magnatone and Bonham history is on www.magnatoneamps.com [4].

In brief, Magnatone's origins started in Los Angeles, California in the late 1930's. Around 1947, Magnatone branding of amplifier and guitar products started and the company name became Magna Electronics.

In the 1950's, Bonham started as a technician with Pacific Mercury Television that made Thomas Organs, progressing to an audio engineer, with Bonham filing two patents for organ circuitry. Bonham's initial vibrato patent # 3,146,292 was filed March 1954 and granted Aug 1964. The interval from '54 to '64 may have resulted from the generality of the patent content, and adverse assessments along the way. Bonham's patent generalises on the specific phase shift circuit technique that Moses described less than a year earlier.

Magna changed ownership to some principals from Pacific Mercury TV, and Bonham came over with the new owners in early 1957 in the role of Chief Engineer, and within a year had introduced his vibrato technique firstly in to the new Custom 200 series of guitar amps.

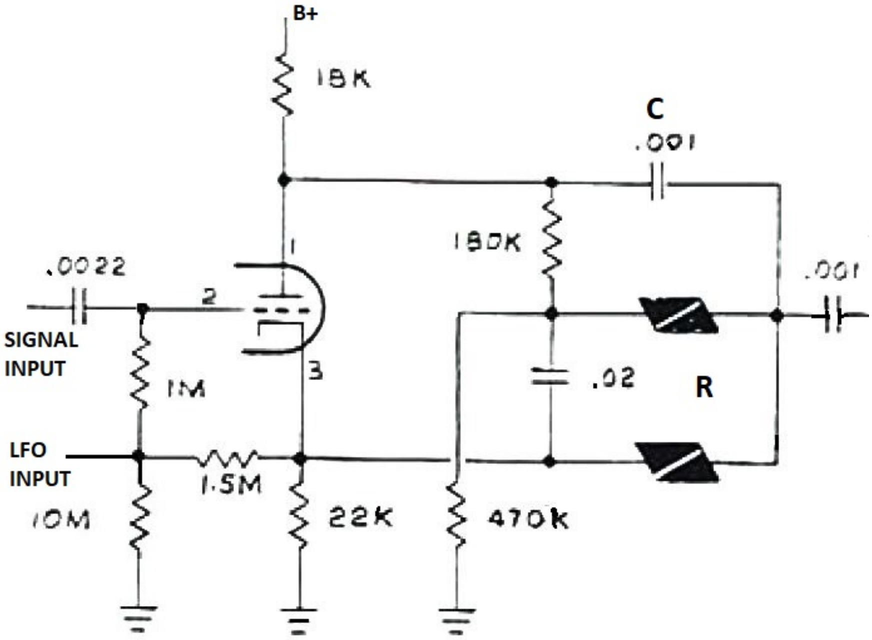
Bonham's patent # 2,988,706 was filed Oct 1958, and granted June 1961, and covers the improved vibrato circuit used in many Magnatone amp models, including the Magnatone 213 assessed on this article.

In 1959, Estey acquired Magna Electronics, who continued to make Magnatone gear, as well as other branded gear. Bonham went on to look at other vibrato effect generation techniques that weren't based on the variable varistor arm in a bridge circuit. Patent # 3,083,606 was filed March 1959, and granted Apr 1962, and a co-patent # 3,160,695 was also filed March 1959, and granted Dec 1964, and those vibrato techniques were used in Magnatone stereo amps. It appears that Bonham had moved on from Magna by mid-1961.

The M2, M4 and 431 models from circa 1964 (now by Estey) use a pared down single stage modulator, as shown in Figure 3, where the signal and the LFO are both mixed at the triode input, and the anode-cathode voltage swing of the triode cathodyne circuit provides both the varistor modulation signal (varistors in series with 180kΩ), and the RC phase modulation signal (where some signal from the 'C' arm leaks over to the 'R' arm via the 180kΩ, and part of the 'R' signal passes through the 0.02uF).

Figure 3

Modulator circuit from M2.



There are several different varistor vibrato circuits employed in Maggie models. Most are just variations of the same basic circuit. Some use a dual stage cascade modulator. By 1967, solid state circuits replaced valve electronics in Magnatone amps, but still used Moses' RC bridge technique for vibrato in the M30 and M35 models, although varistors were now replaced by an incandescent lamp and LDRs.

Varistor Information

The varistor that Magnatone used was made from Silicon Carbide (SiC) using a ceramic manufacturing technique. SiC varistors were being widely used in telephone sets in the 1950s, and also for over-voltage clamping applications in electrical equipment [App.C].

Varistors for this application are specified at low currents, typically 1mA or less, and high-ish voltages of typically 50-120V. A varistor manufacturer's datasheet shows a characteristic V-I (voltage versus current) curve for a part, and the Carborundum 233BNR family curves are shown below. Apart from a non-descript model number, varistors are usually identified at a particular V-I point to assist differentiation and application selection, given that the device is a non-linear resistance.

Magnatone amps appear to have been designed for a varistor with a curve that is close to the following nominal V-I points: 0.01mA @ 32V; 0.05mA @ 55V; 0.1mA @ 65V; 1mA @ 105V.

It is understood that the Kanthal Globar (previously Ceslwld & Carborundum) 233BNR-32 was the model used by Magnatone, with the Workman FS1203 and FS605 being equivalent replacement parts [5]. Service outlets and part suppliers often had a large range of Workman and Zenith varistors for many applications. A few such parts have a V-I rating that is close to the 233BNR-32. Metrosil is the only manufacturer presently producing SiC varistors [6].

Workman FS1203: 0.05 mA @ 49V (Zenith 63-4906)

Workman FS605

Workman FS1211: 0.05 mA @ 61V (Zenith 63-5327)

Workman FS1205: 0.05 mA @ 80V (Zenith 63-5058)

Workman FR1039: 1.0 ma @ 68V

Workman FS927: 1.0 ma @ 80V

Workman FS308: 1.0 Ma @ 110V

Metrosil 100-P/W/921: 0.1 mA @ 55V; 1.0 mA @ 100V.

Figure 4.

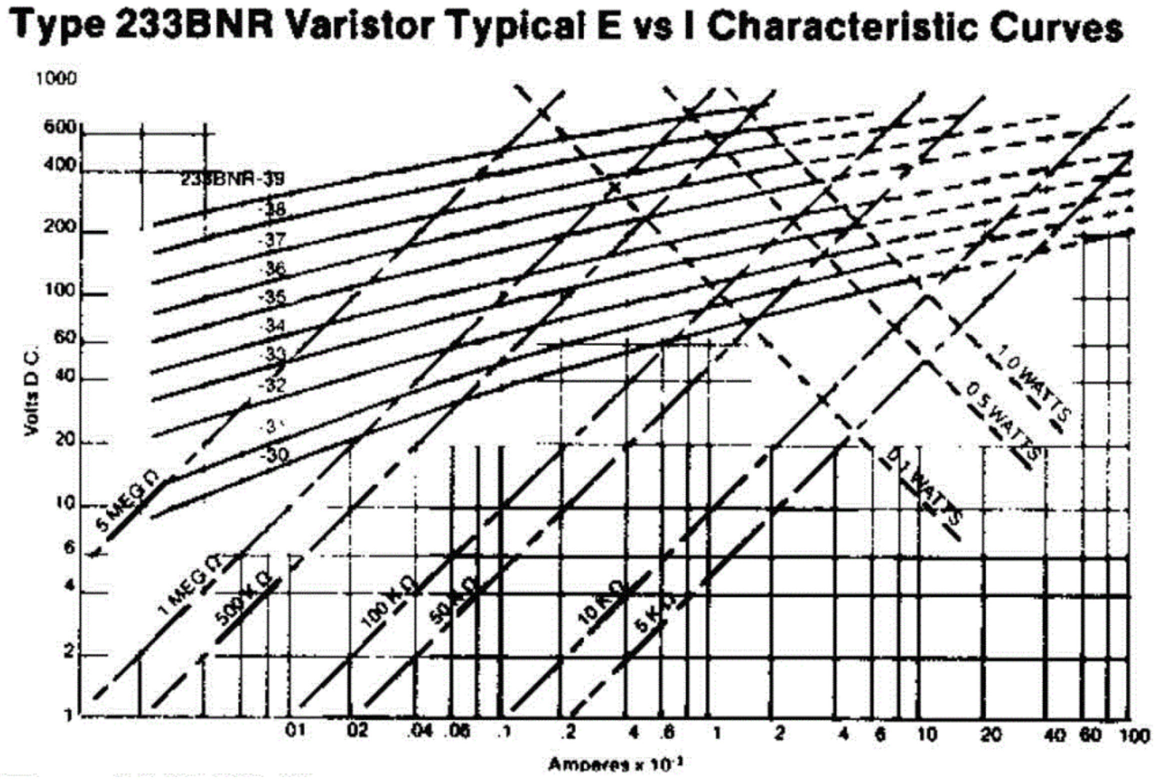
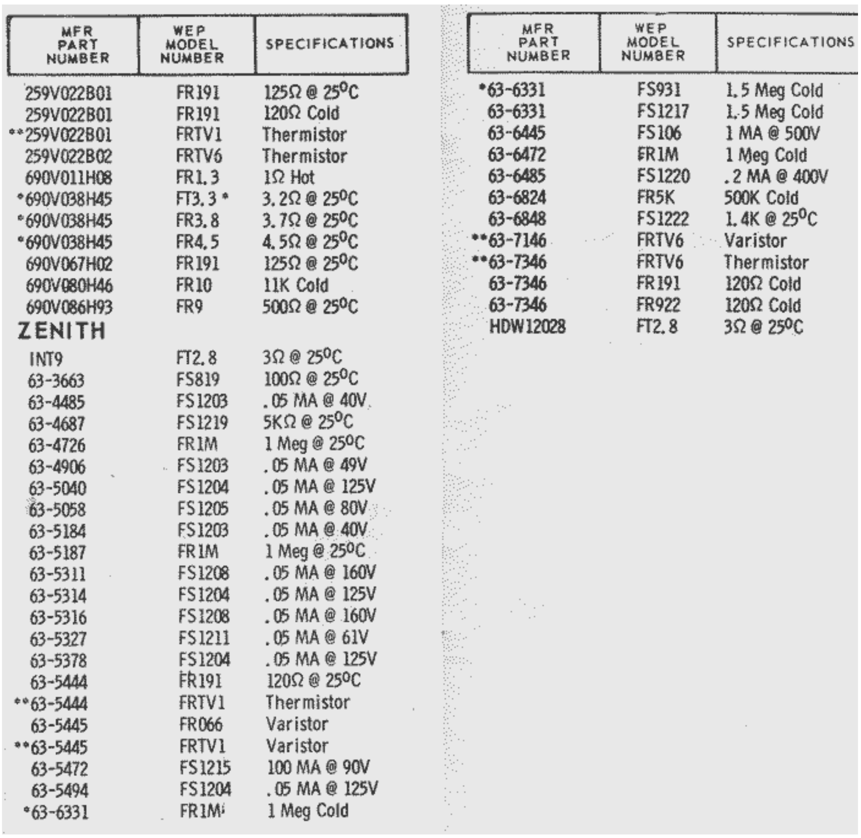
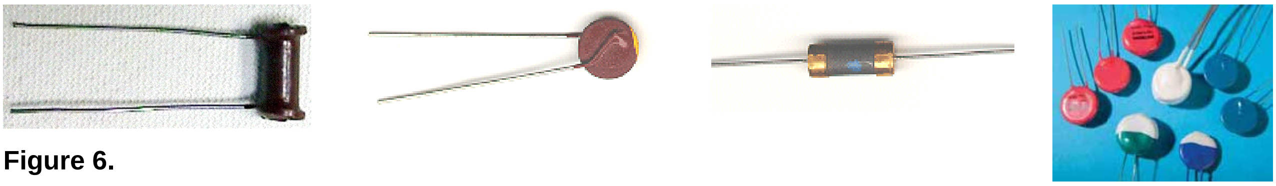


Figure 5.



MFR PART NUMBER	WEP MODEL NUMBER	SPECIFICATIONS
259V022B01	FR191	125Ω @ 25⁰C
259V022B01	FR191	120Ω Cold
**259V022B01	FRTV1	Thermistor
259V022B02	FRTV6	Thermistor
690V011H08	FR1.3	1Ω Hot
*690V038H45	FT3.3 *	3.2Ω @ 25⁰C
*690V038H45	FR3.8	3.7Ω @ 25⁰C
*690V038H45	FR4.5	4.5Ω @ 25⁰C
690V067H02	FR191	125Ω @ 25⁰C
690V080H46	FR10	11K Cold
690V086H93	FR9	500Ω @ 25⁰C
ZENITH		
INT9	FT2.8	3Ω @ 25⁰C
63-3663	FS819	100Ω @ 25⁰C
63-4485	FS1203	.05 MA @ 40V
63-4687	FS1219	5KΩ @ 25⁰C
63-4726	FR1M	1 Meg @ 25⁰C
63-4906	FS1203	.05 MA @ 49V
63-5040	FS1204	.05 MA @ 125V
63-5058	FS1205	.05 MA @ 80V
63-5184	FS1203	.05 MA @ 40V
63-5187	FR1M	1 Meg @ 25⁰C
63-5311	FS1208	.05 MA @ 160V
63-5314	FS1204	.05 MA @ 125V
63-5316	FS1208	.05 MA @ 160V
63-5327	FS1211	.05 MA @ 61V
63-5378	FS1204	.05 MA @ 125V
63-5444	FR191	120Ω @ 25⁰C
**63-5444	FRTV1	Thermistor
63-5445	FR066	Varistor
**63-5445	FRTV1	Varistor
63-5472	FS1215	100 MA @ 90V
63-5494	FS1204	.05 MA @ 125V
*63-6331	FR1M	1 Meg Cold
*63-6331	FS931	1.5 Meg Cold
63-6331	FS1217	1.5 Meg Cold
63-6445	FS106	1 MA @ 500V
63-6472	FR1M	1 Meg Cold
63-6485	FS1220	.2 MA @ 400V
63-6824	FR5K	500K Cold
63-6848	FS1222	1.4K @ 25⁰C
**63-7146	FRTV6	Varistor
**63-7346	FRTV6	Thermistor
63-7346	FR191	120Ω Cold
63-7346	FR922	120Ω Cold
HDW12028	FT2.8	3Ω @ 25⁰C



Original maggie varistor

FR1039 : 1.0 ma @ 68V

FS-308 : 1.0 Ma @ 110V

Metrosil range

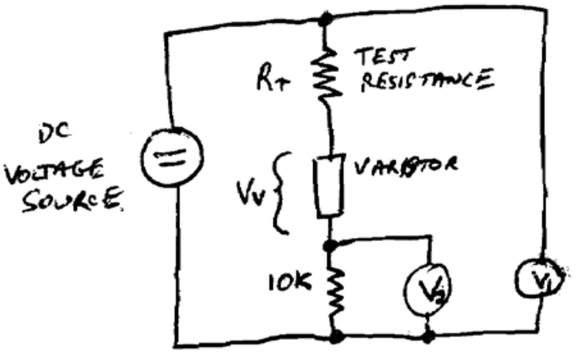
Figure 6.

Varistor V-I measurement

Varistors can't really be measured by a multimeter on its resistance setting. As indicated in the previous section, a spot measurement of V-I is applicable. For this, a variable DC supply of 50 to 100V would be appropriate, and a test circuit would use a current sense resistance (eg. 10kΩ) in series with the varistor. Alternatively, a modern insulation resistance meter with a 100Vdc range would show a resistance of 100kΩ at 1mA.

To measure a characteristic curve, a higher DC voltage supply can be used with a range of additional test resistors that are placed in series with the varistor under test. I have used a 243VDC supply with a range of test resistors from 220kΩ to 24MΩ. Most test meters won't significantly influence the measurement.

Measured varistor V-I levels for the same part may show at least +/-10% differences. Matching pairs of SiC varistors to put in a vibrato circuit is worthwhile to ensure the parts are nominally the same, but it is unlikely that any noticeable advantage is gained from close matching.



$$V_V = V_1 - (R_T + 10k)^{*} \frac{V_2}{10k}$$

$$I_V = \frac{V_2}{10k}$$

Figure 7. Varistor V-I test circuit.

Cloning a Magnatone varistor

Cloning a Magnatone varistor has been attempted in two ways using Zener diode and MOV parts to emulate regions of the varistor V-I curve. A clone can better represent the original varistor non-linear V-I curve as more clone parts are used to emulate smaller regions of the V-I curve.

A long series string of low voltage (<5V) zeners, along with a few resistors to trim the characteristic to match an original varistor, has shown close performance, both in V-I characteristic curve and in spectrum analysis of a signal passed through the Magnatone circuit. Low voltage zeners exhibit a softer resistance transition as the applied voltage approaches the zener breakdown voltage than higher voltage zeners (> 5V1). This clone is ‘polarised’, as it is effectively a zener diode, and so needs to be correctly oriented in the Magnatone circuit.

Martin Manning [7] has prepared a simpler parallel configuration using three MOVs, each with a trimming resistor. A MOV V-I curve doesn’t match that of a SiC varistor, but it is a closer match than a high voltage zener, and MOVs of suitable voltage rating (8V to 100V) are easily purchased. This clone is not polarised, so is easier to use.

Limited in-situ amp testing with this clone so far indicates good equivalent performance to the original varistors.

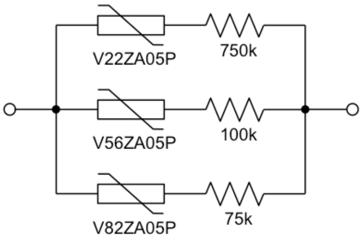


Figure 8.

SiC Varistor Substitute Using MOV’s.
Suggested part numbers Littelfuse ZA series

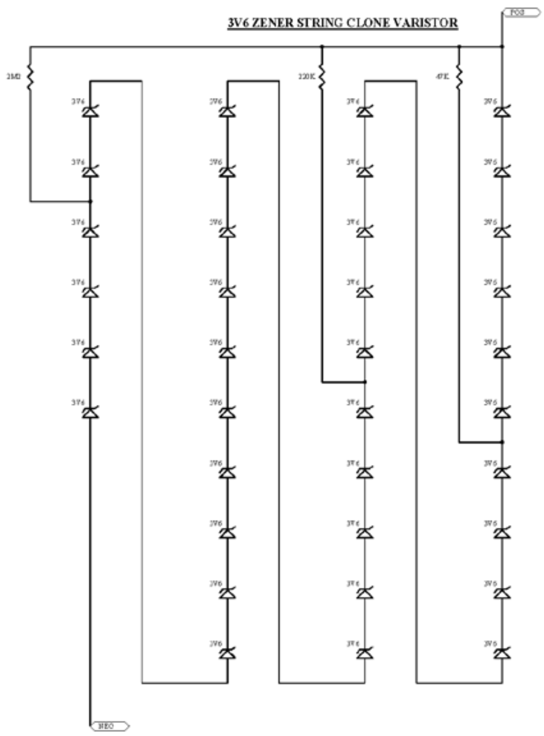
Some results are presented below that relate to two original SiC varistors, a 3V6 zener string clone, and an early version of Martin Manning’s simpler clone that used zeners.

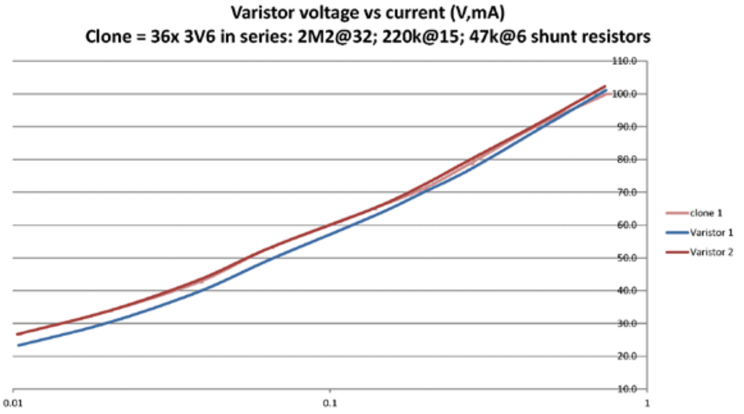
Spectrum responses of a 1kHz signal through a Magnatone vibrato circuit are presented in Figure 10. Note the harmonics of the ~10Hz LFO signal, the residual 50 and 100Hz mains ripple, the 100Hz modulation sidebands of the 1kHz signal, and the second and third harmonics of the 1kHz signal. Also note that this vibrato effect adds discrete LFO sidebands to all the other signals, rather than a vibrato technique that would vary just the 1kHz as a continuous spectrum with a small bandwidth.

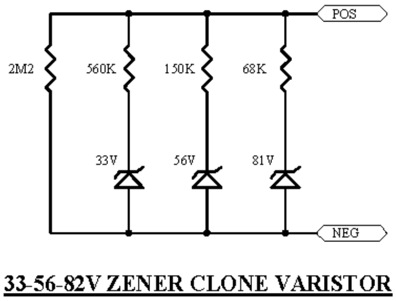
Two spectrum plots are shown: the first plot covers the audio spectrum with one form of FFT window using the TrueRTA application; the second plot is an octave either side of the 1kHz signal, using the REW5 application with a different FFT window.

Only limited in-situ amp testing has been done with these two clones, with no adverse performance.

Figure 9. V-I characteristic curves of clone versus two original Magnatone varistors.







The V-I curves are MS Excel smoothed curves of 9 measurement points.

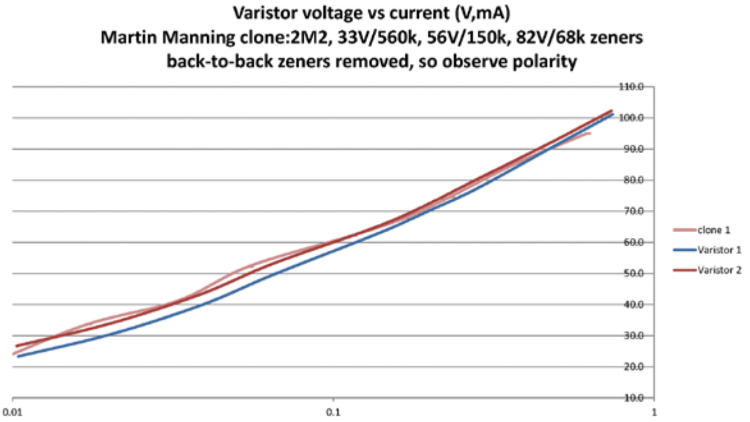
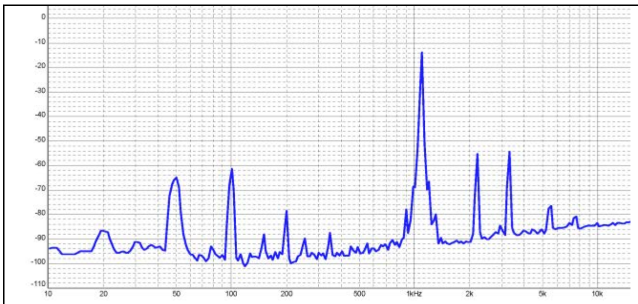
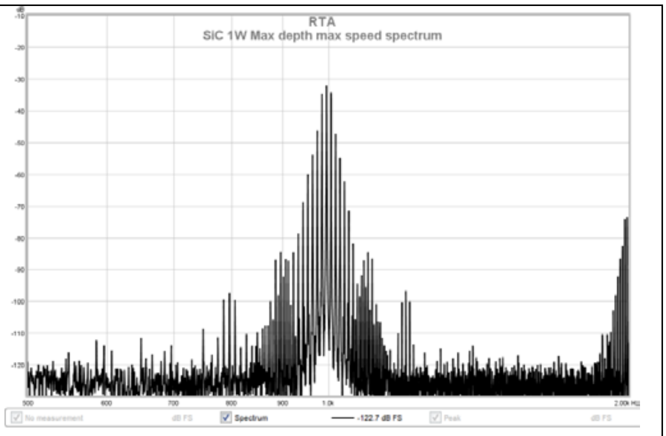
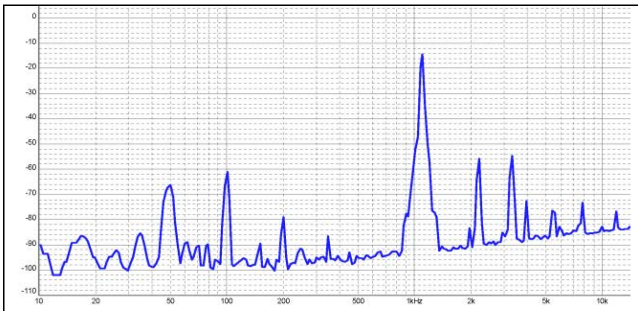


Figure 10. Spectrum responses clone versus original varistors in Magnatone 213 varistor circuit.

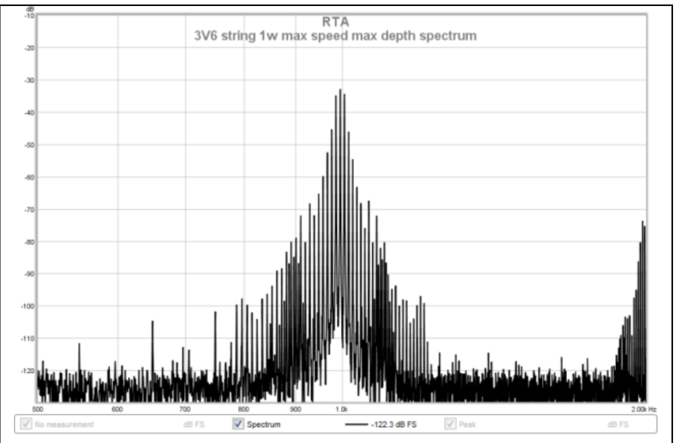


Original Magnatone varistor. 100Hz mains ripple sidebands noticeable around 1kHz signal. Steep fall-off of LFO sidebands.

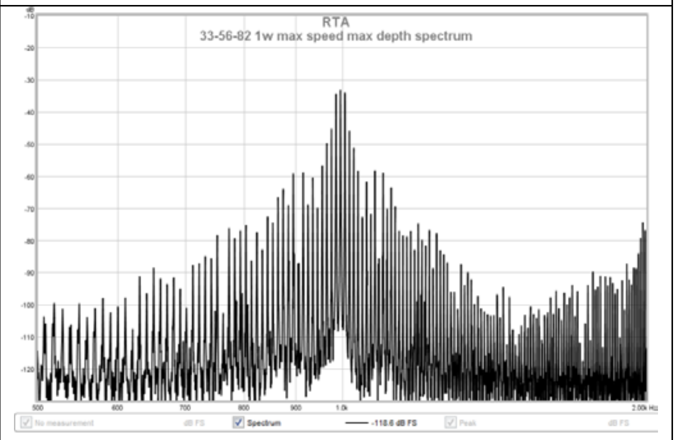




3V6 zener string clone varistor. Higher levels of LFO harmonics. Broader LFO sidebands around 1kHz.



33/56/82V zener clone varistor. High levels of higher order LFO harmonics. Much broader LFO sidebands around 1kHz.



Recent testing (in 2021) used available Littlefuse ZA series 5mm disk MOVs with V08, V39 and V82 part numbers, and better instrumentation. Just using a parallel combination of V08 +1M6 // V39 + 330k // V82 + 33k gave a good V-I match across most of the 10V to 120V range, but with a noticeable variation around 70V. MOV availability didn't allow other voltage parts to be used, so the match was improved by also including two V39 MOVs in series to make a clone with V08 +1M6 // V39 + 330k // 2x V39 + 100k // V82 + 68k. The V-I comparison shows how each MOV aligns with a section of the original varistor curve, but also shows the minor deviations at the overlap of when each MOV contributes to the curve.

Testing of the MOV-based clone was done in a clone amp with Magnatone vibrato type circuitry. Only the signal modulator stage of circuitry was tested, with no output loading on the modulator stage except for a 10MΩ load and a 100x scope probe. In comparison to the original SiC varistors, the MOV cloned varistors showed the following differences:

- The LFO signal leakage at the output of the modulator (with no input signal to the modulator, and maximum Depth setting) was noticeably higher.
- With signal input to the modulator, and maximum Depth setting, the LFO sideband levels of the signal were noticeably lower.
- With minimum Depth, a frequency response of the signal modulator stage showed a -7dB dip centred around 2-3kHz, and a related changed phase response above about 600Hz.
- The LFO modulated signal envelope at maximum Depth was noticeably jagged and distorted.

A likely reason for some of the noticeable differences is poor symmetry of the clone V-I curve about the idle point, due to the MOV voltage regions not being symmetrically aligned with the idle point. The idle voltage of a varistor is about 55V for a 250V supply rail to the LFO modulator, with a varistor current of nearly 0.1mA. The V39 MOV part measured 35V for 0.1mA, and so 2x V39 parts would drop 70V, which are on either side of a 55V idle level. Another likely reason is the V-I deviation when operating between MOV voltage levels.

Perhaps a clone with a smoother V-I response would include say five MOV voltages like V08, V33, V47, V68, and V100, although part availability can become an issue. The region around idle may be better emulated by a blend of V47 and V68 MOVs, or similar voltage parts that operate closer to 55V at 0.1mA.

Circuit Design – Magnatone 213

The Magnatone 213 circuit is shown below. After the input 12AX7 amplifier stage, and the volume and tone controls, the instrument signal is amplified by a 12AX7 triode and DC coupled to the vibrato stage signal phase modulator – a 12AU7 triode cathodyne (or split-load) stage with uneven load resistances of 18kΩ to power supply ‘A’, and 82kΩ to 0V.

The resistance-reactance bridge arms are formed by the 820pF ‘reactance’ arm connected to the 12AU7 anode, and the ‘resistance’ arm connected to the 12AU7 cathode. The variable resistance arm effectively comprises the two varistors in parallel, where the 47nF coupling capacitor in ‘series’ with each varistor allows the signal to pass, and allows the varistor resistance to be modulated by the other 12AU7 triode acting as a Low Frequency Oscillator (LFO) modulator. The LFO modulator 12AU7 triode is in a cathodyne circuit with 3k9Ω cathode bias and 2M2Ω grid leak, and a load resistance of 47kΩ to ‘A’, and 43k+3.9k ≈ 47kΩ to 0V.

The LFO is a simple RC phase shift oscillator using both 12AX7 triodes in parallel, with a rate pot, an output signal level pot, a remote pedal disable switch, and RC low pass filtering to the LFO modulator.

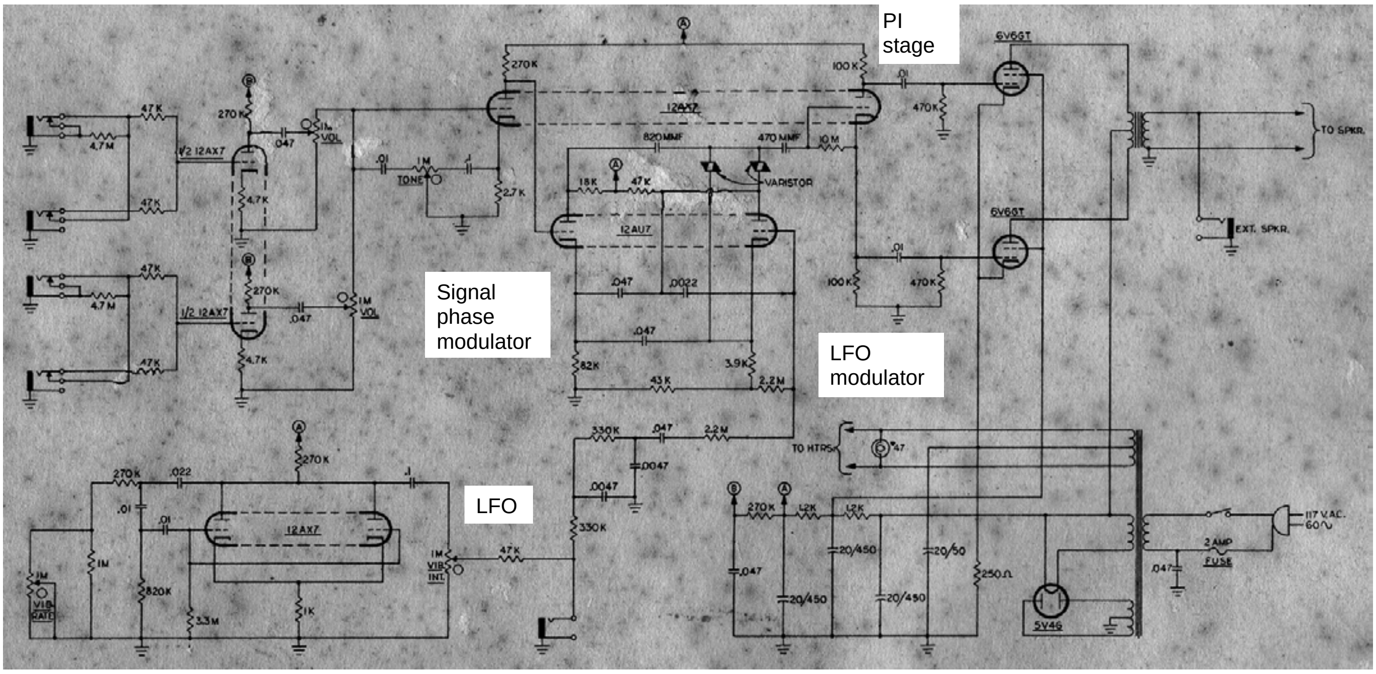


Figure 11. Magnatone 213 circuit.

The combined output of the resistance-reactance bridge arms is then capacitor coupled via 470pF to a grid-leak biased cathodyne PI stage driving the 6V6 pentode push-pull output stage. The PI stage presents a very high load impedance on the phase modulator output, which is needs to avoid excessive tremolo occurring.

The Output stage HT supply voltage is about 350V at idle. Supply voltage at ‘A’ is estimated to be about 300V, with probably 10mA through the 1k2 dropper to ‘A’, and 10+15= 25mA through the 1k2 dropper from the HT supply.

Signal phase modulator

A 12AU7 loadline is shown below for a 300V supply, and 82k + 18k = 100kΩ load. The previous 12AX7 triode loadline (see below) indicates that the 12AU7 grid is about 155V. With about 155+6 ~ 160V across the 82kΩ cathode resistor, the cathodyne current is about 160/82k = 2.0mA, with grid bias of about -6V, and an anode-cathode voltage of about 120V.

Full signal voltage swing across the 12AU7 triode is generated from about 6Vpk input swing, which would generate about 2.3-1.5 = 0.8mApk anode signal swing. The anode voltage swing would then be about 0.8x18 ~ 14Vpk, and the cathode voltage swing about 0.8x82 ~ 65Vpk. Maximum design level heater-cathode voltage is 200V, so this stage is exceeding that limit at peak signal voltage of 155+64 ~ 220V.

Each 47nF coupling capacitor, connecting to a varistor end, introduces a CR high pass filter to the resistance arm input signal, where R is approximately the connected 47kΩ resistance connected to ‘A’, or 0V. The filter corner frequency is about 70Hz, so is not really noticeable for a guitar.

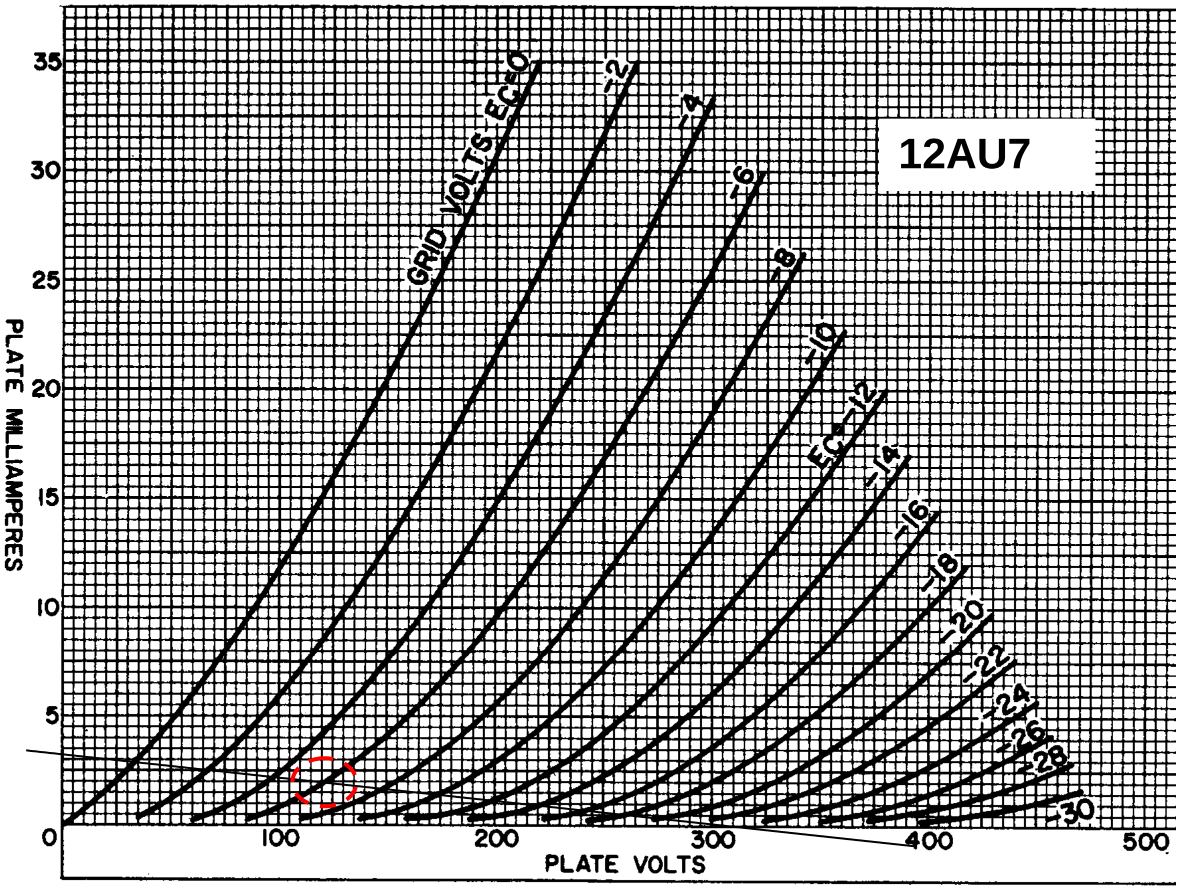


Figure 12. Signal modulator 12AU7 triode loadline.

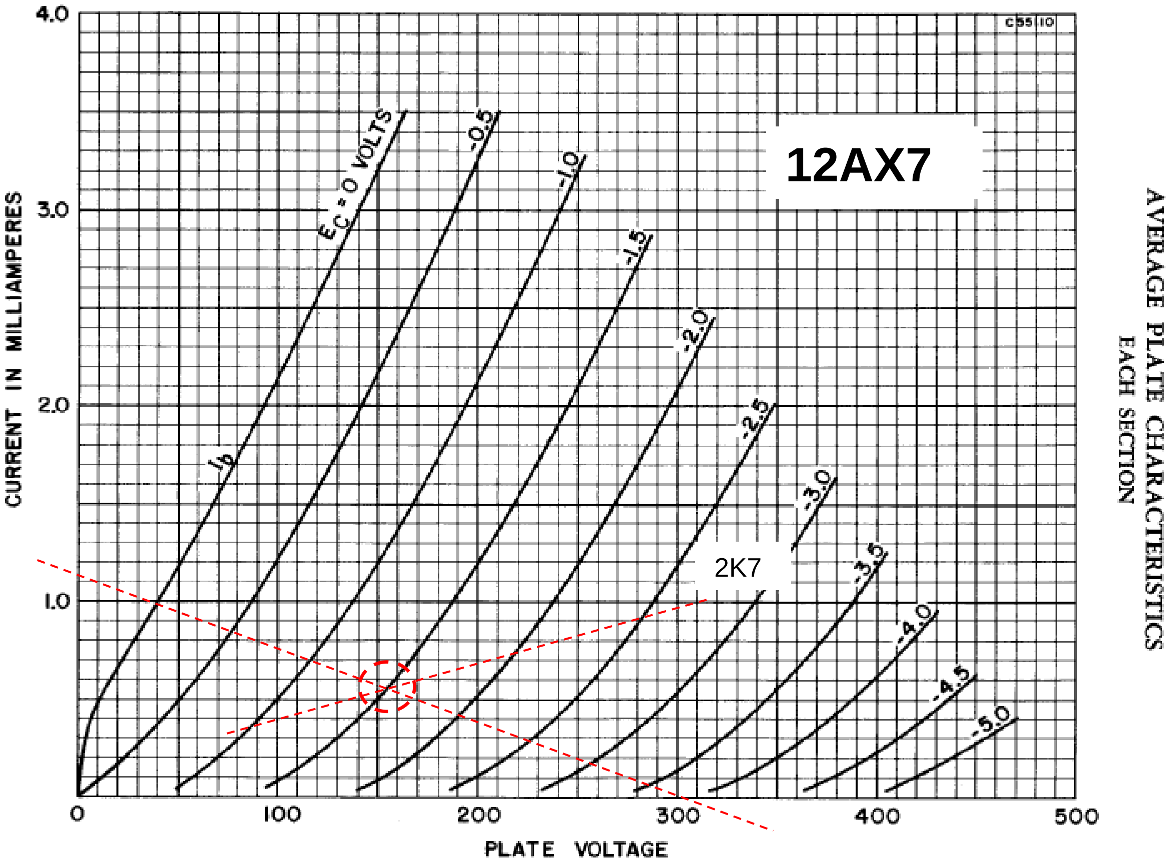


Figure 13. 12AX7 triode driving the signal modulator.

LFO modulator

A 12AU7 loadline is shown below for a 300V supply, and 47k + 47k = 94kΩ load. With 3k9 cathode bias, the anode current is about 2mA, with about 120-130V between anode and cathode. However, the series varistor connection provides a bypass current around the triode. With an anode-cathode voltage of 120-130V, and with ~0.1mA passing through the varistors, then the bias point hardly moves at all (the anode-cathode voltage would drop a bit as less current is flowing through the triode).

When the LFO modulator triode is driven towards saturation, the voltage across each varistor decreases and hence varistor resistance increases and hence bypasses less current around the triode. Similarly, when the LFO modulator triode is driven towards cut-off, the voltage across each varistor increases and hence varistor resistance falls and the varistors bypass more current around the triode.

With the 12AU7 driven to saturation, the anode-cathode voltage could reduce to about 40V, ie. about 20V across each varistor. The varistor resistance is then very high (>1MΩ), and has negligible effect on the triode loadline.

With the 12AU7 driven towards cut-off, the anode-cathode voltage increases towards the supply voltage and the triode resistance increases. However, the varistor resistance is falling, and effectively determines the max anode-cathode voltage level due to the 47kΩ voltage drops. The equivalent circuit then tends towards a 300V supply with 94kΩ fixed resistance in series with two varistors, where each varistor operates at about 100V and 1mA, with the 94kΩ dropping about 100V. The result is the triode loadline drooping to about a 200V x-axis crossing.

The effect on the triode loadline is quite non-linear, and only really becomes noticeable for triode voltage above the idle level of about 120-130V.

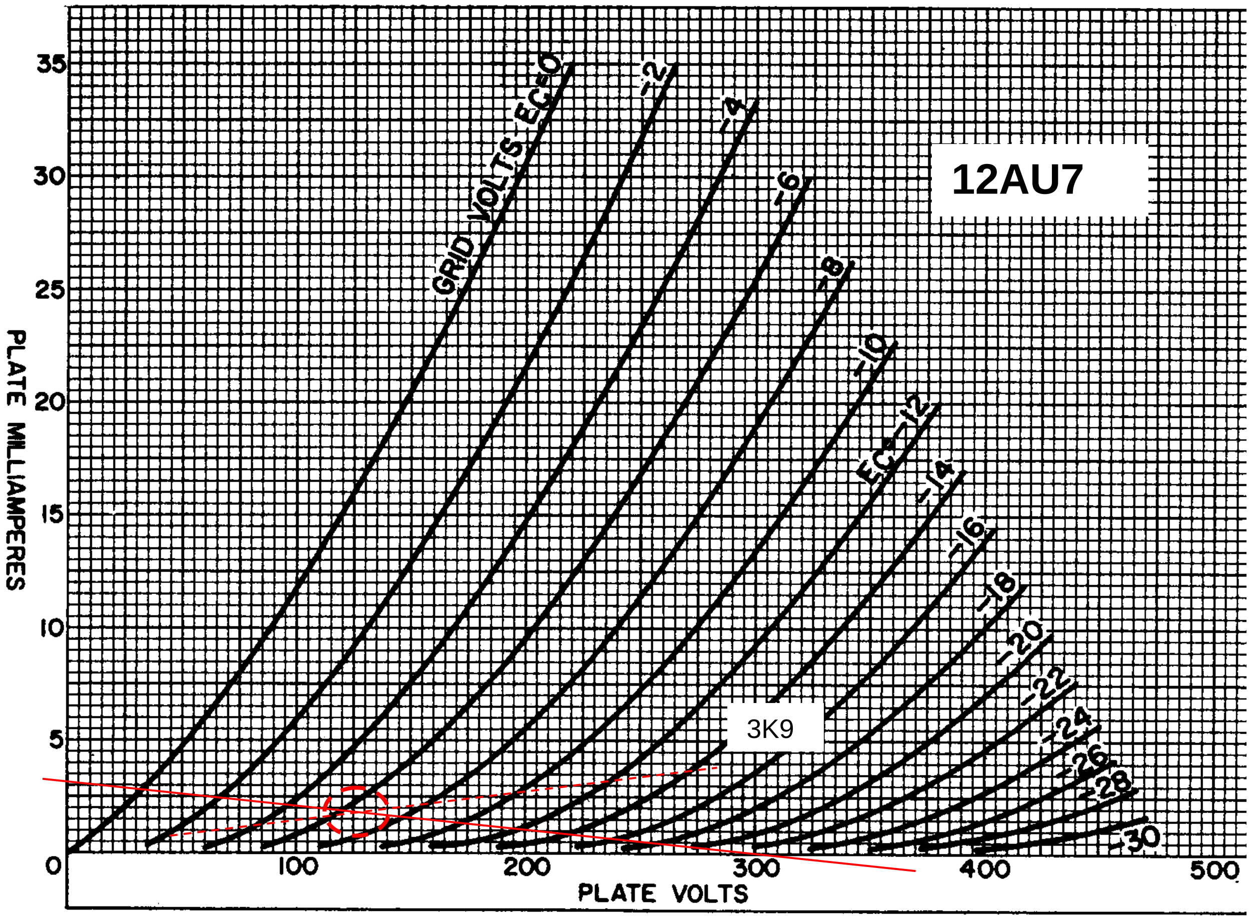


Figure 14. LFO modulator 12AU7 triode loadline.

Assuming signal frequencies where the 47nF coupling caps are insignificant, and using Moses' representation (initially assuming the electrical centre of the AC source is the mid-point of the 12AU7), then the resistance arm is the varistor resistances in parallel (each varistor initially assumed to be 55V/0.1mA = 550kΩ each). The RC time constant of 225kΩ and 820pF is about 180 Ω.uF, which achieves about 90deg output phase lag at 1kHz from the two arms.

At min varistor resistance of about 100V/1mA = 100kΩ each, the RC time constant of 50kΩ and 820pF is about 41ΩuF, which achieves about 150deg output phase lag at 1kHz from the two arms. At max varistor resistance of about 25V/0.01mA = 2.5MΩ each, the RC time constant of 1.2MΩ and 820pF is about 1000ΩuF, which achieves ~10deg output phase lag at 1kHz from the two arms.

Maximum design level heater-cathode voltage is 200V, so this stage likely exceeds that limit at cathode peak of about 130 +80 = 230V.

It is likely that varistors are somewhat hygroscopic, so long-term environmental conditions may modify the max varistor resistance achieved at very low currents. Any parasitic capacitance across each varistor would reduce the phase shift affect at higher frequencies.

A key aspect of the LFO modulator is that the midpoint of the series varistors stays effectively at a stable DC voltage, with little LFO signal to leak through to the output stage and speaker as LFO thumping. Any residual LFO signal would be attenuated by the 10nF-470kΩ high pass filter (~33Hz corner frequency) to each 6V6, and then by the push-pull symmetry.

Low frequency oscillator

The LFO is a single amplifier stage (itself providing 180 degrees of phase shift) with a three CR feedback phase-shift network (providing a further 180 degrees phase shift, = 360 degrees total shift). The total loop gain is made greater than unity using an amplifier valve with effective gain >29. A fairly large 270kΩ anode resistor is used with the 12AX7, to maximize gain and output swing. 14 amp models use simple 3.3MΩ grid leak biasing, with cathode tied to ground, where the grid dc level is pumped up during operation from grid capacitor signal input. The 213, 440, 480 models also include a 1kΩ cathode bias resistor.

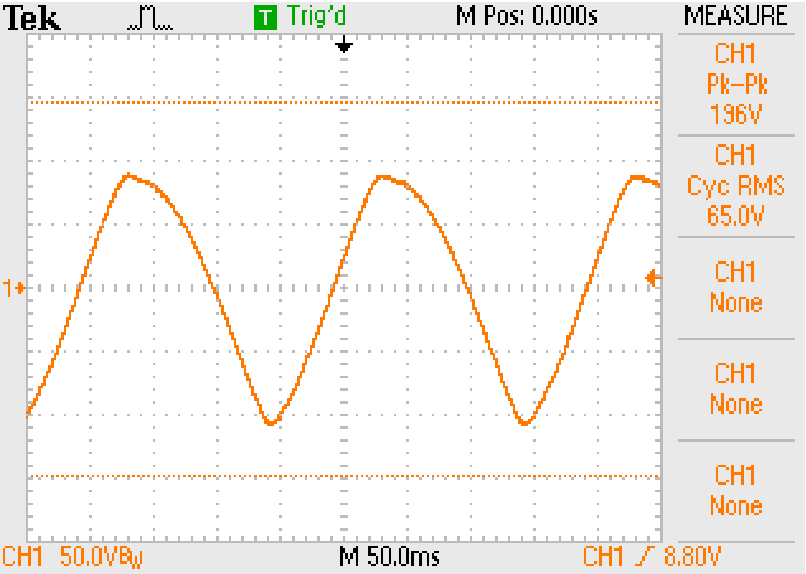
The 213 model LFO with 300V supply generates an output signal level of about 69Vrms with 12.6% THD, however this is the only Magnatone model using paralleled triode stages.

The 14 amp models using simple 3.3MΩ grid leak biasing have a slightly lower output signal level of about 66Vrms with 13.0% THD. The other 2 amp models with the 1kΩ cathode bias resistor (but only one triode) have a slightly lower output voltage of 65Vrms, with a lower THD of 11.5%, as shown in Figure 15 and Figure 16. The unbypassed 1kΩ lowers the triode gain to achieve a lower THD but doesn't reduce gain to a level where oscillation fails to start. Figure 17 indicates the triode loadline and likely nominal operating point.

One simple method to suppress harmonic levels is to connect a capacitor between anode and grid, which increases feedback as frequency increases. For the example 5Hz waveform, adding 100pF lowered THD to 9.1%, with 100Hz harmonic level decreasing by 25dB, although other operational issues may indicate that something like 47-68pF may be a practical upper limit. Other methods such as increasing the grid leak resistance, and shunting the speed pot with capacitance can similarly lower THD, but with the disadvantage of also lowering signal output level.

Figure 15.

LFO waveform.



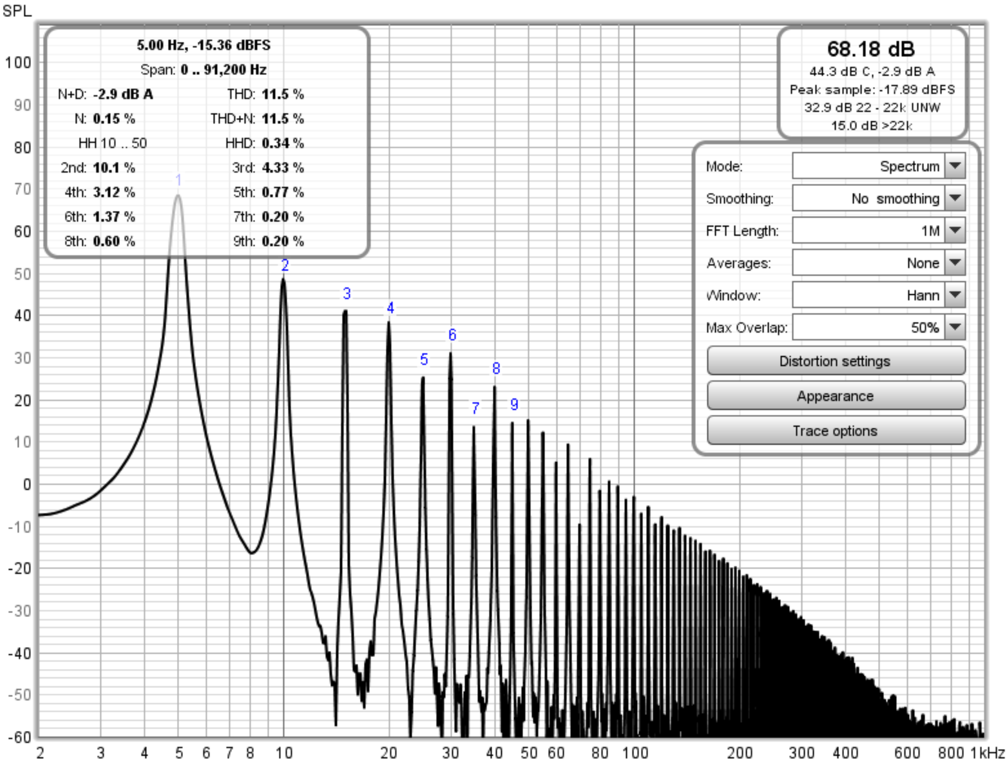


Figure 16.

LFO spectrum.

5.00 Hz, -15.36 dBFS
Span: 0 .. 91,200 Hz

N+D:	-2.9 dB A	THD:	11.5 %
N:	0.15 %	THD+N:	11.5 %
HH 10 .. 50		HHD:	0.34 %
2nd:	10.1 %	3rd:	4.33 %
4th:	3.12 %	5th:	0.77 %
6th:	1.37 %	7th:	0.20 %
8th:	0.60 %	9th:	0.20 %

68.18 dB
44.3 dB C, -2.9 dB A
Peak sample: -17.89 dBFS
32.9 dB 22 - 22k UNW
15.0 dB >22k

Mode: Spectrum
Smoothing: No smoothing
FFT Length: 1M
Averages: None
Window: Hann
Max Overlap: 50%

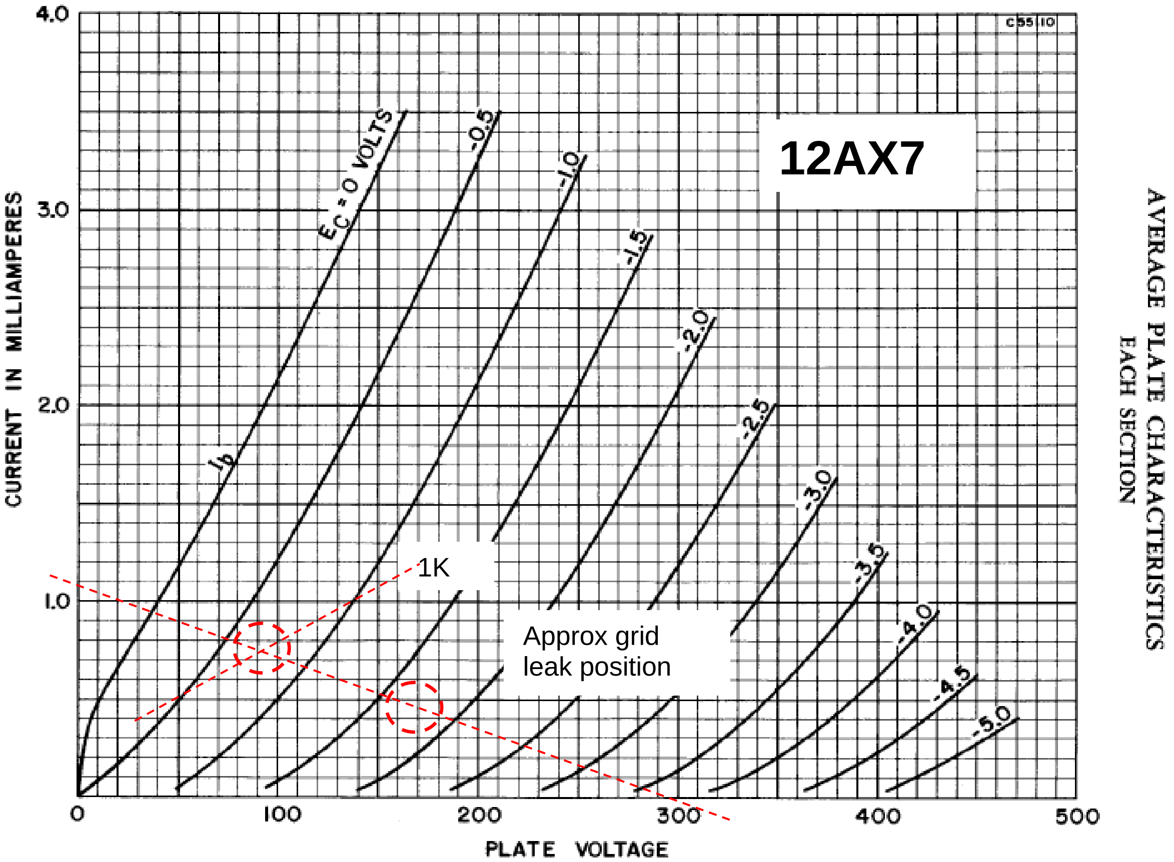


Figure 17. LFO 12AX7 single triode loadline.

Although the use of fixed bias from a LED allows simple indication of the LFO working, a 1.73V amber LED generated 15.1% THD with a 64.5Vrms signal output.

Two RC filters (380k/4n7 and 330k/4n7) with 100Hz corner frequency are used to attenuate signal band harmonics without significantly attenuating the LFO fundamental or its lower order harmonics.

Amplitude modulation

Part of the perceived advantage of the Magnatone vibrato effect appears to be that it also introduces some tremolo amplitude modulation along with the frequency modulation. The plot on the right is the amplifier output signal with the vibrato circuit using standard varistors. The plot time-frame shows the vibrato frequency, with the 1kHz signal showing as the solid part.

The plot is when LFO depth pot is at maximum setting in this particular amp. At lower pot settings, the amplitude modulation is much more sinusoidal.

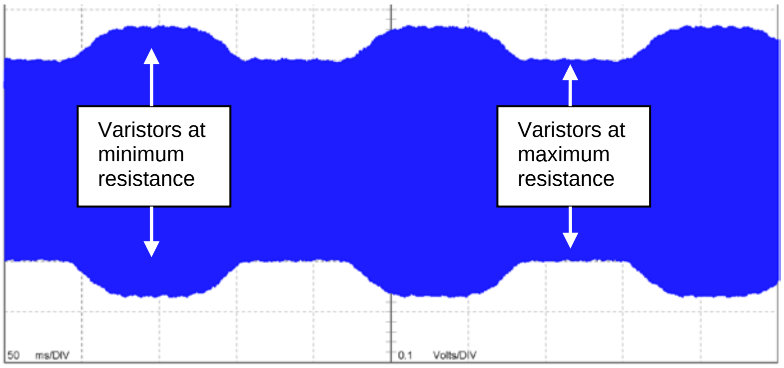


Figure 18. 1kHz tone signal through Magnatone circuit.

The amplitude modulated signal is at a maximum level when the varistor resistance is at its lowest value due to the summing of voltages from the C and R arms of the bridge circuit. The flat topping effect indicates that the limit of effective phase shift has been reached. In Figure 1, the phase shift variation reaches a practical limit when R/X = ωRC becomes too small, or too large. This may be why Magnatone introduced some amp models with two sequential stages of phase shift, so as to extend the vibrato effect with less amplitude non-linearity.

The amplitude harmonic structure arising from the vibrato phase modulator stage, and the subsequent low frequency response of the amplifier itself and the speaker system, will determine how much the tremelo effect is noticeable. Excessive phase shift modulation, such as the ‘square-wave like’ tremelo shown in Figure 18 would increase the higher order harmonic content. Any imbalance of the two varistor arm resistances would also add in some tremelo contribution to the output of the phase modulator stage.

The tremelo amplitude varies with the LFO signal as set by the vibrato intensity pot, followed by the high-pass filter response of the modulator RC arms. As such, it could plausibly be neutralised by an active gain and single pole low pass filter circuit.

For Figure 18, the test amp accentuates the tremolo effect as the phase modulator is loaded by a 0.5MΩ pot rather than the very high impedance of the typical Magnatone PI stage. If the amplitude modulation signal is clipping too much at the minimum or maximum levels, then the LFO modulator can be biased hotter or colder respectively by lowering or raising the value of the 3k9Ω cathode bias resistor.

The modulation characteristic changes with frequency, so the results of 1kHz are somewhat in the middle. This characteristic also shows the benefit in using varistors for the variable arm, as the resistance can be made to change over about a 25:1 ratio, which is a very substantial non-linear range.

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Acknowledgements

Great assistance was provided by Ken Stone and Charles Wood for Schober organ details and circuits, and Stephen Keller and Martin Manning for varistor and technical advice. Dinko Tomljanovic provided NOS varistors and Gary Croteau provided varistor details.

Appendix A. Schober vibrato circuits using varistors

Circa 1960, Schober's Concert and Consolette models, and the new smaller Spinet model, used varistors in circuitry similar to Magnatone's. In Dec 1961, a letter response by Dorf identifies the new Spinet model, and an advert in RTV&H shows off the Spinet and describes recent changes to the Consolette. A letter in June 1962 RTV&H describes the Schober vibrato circuit as using varistors. But by 1963 the varistor-based vibrato circuits were replaced by transistorised circuitry using LDR variable resistances.

Taken from Schober's organ construction booklet, the introduction dates of the PR-2 and PSR-2A pre-amplifier – vibrato pcb assemblies shown below are not known, although a noticeable progression in circuit refinement can be gleaned from the 'A' revision. A PSR-2 schematic is not shown, but is very similar to the PR-2 schematic, with changes just to suit the organ model. The construction booklet identifies the varistor as BNR type Globar.

For the PR-2 circuit, B+ is about 410V, with X about 390V and the signal modulator stages V2A, V2B idle at about 9mA each. Unlike Bonham's circuitry, the LFO signal from V3B is capacitor coupled to the varistors, and the varistor is DC biased via the BIAS pot and 1MΩ resistors 20 and 21. There also seems to be an error in the value of resistor '31'.

Each varistor appears to idle at about 55-60V, with about 65uA, so a common varistor model rating like 60V @ 0.05mA (eg. Carborundum 233B NR-32), or 40V at 0.05mA would seem appropriate. Given the 1nF modulator capacitor arm is very similar to Magnatone's 820pF, the Schober modulator middle frequency is likely to also be near 1kHz.

The varistor LFO voltage swing could be between about 20 to 100V, with peak current up to nearly 1mA. The LFO PI can source a peak LFO signal current via the varistors of about (350-120)V/(18k+100k+100k+18k) = 1mA. Schober's circuit provides a nice high impedance to the varistor centre point summing junction due to the valve grid loading.

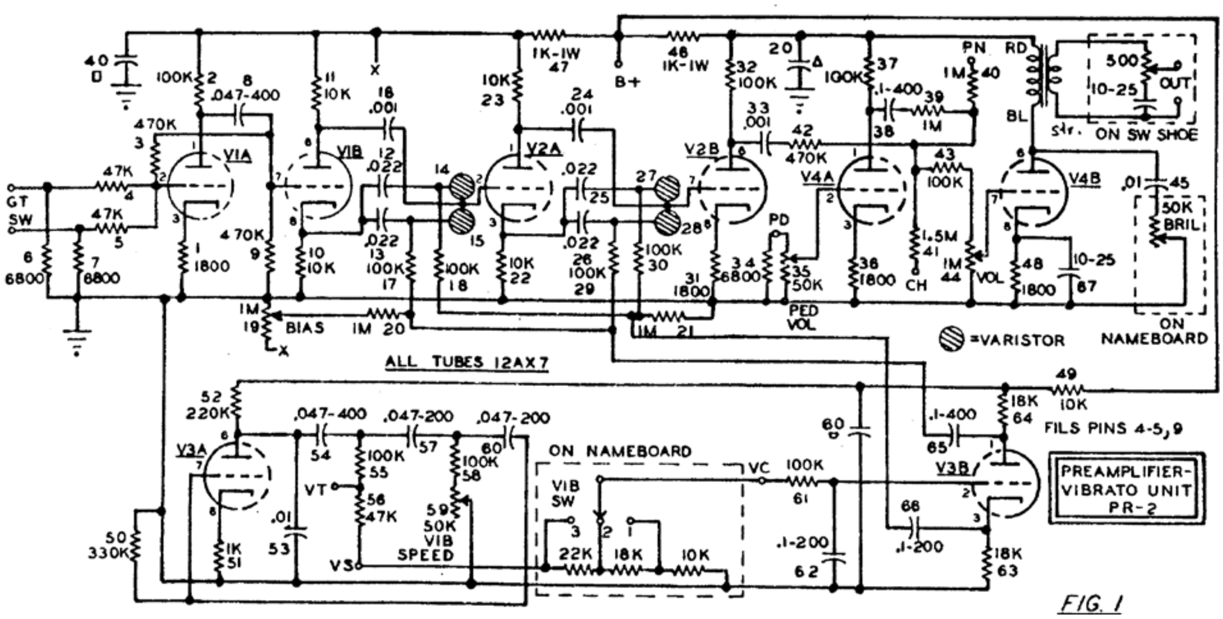
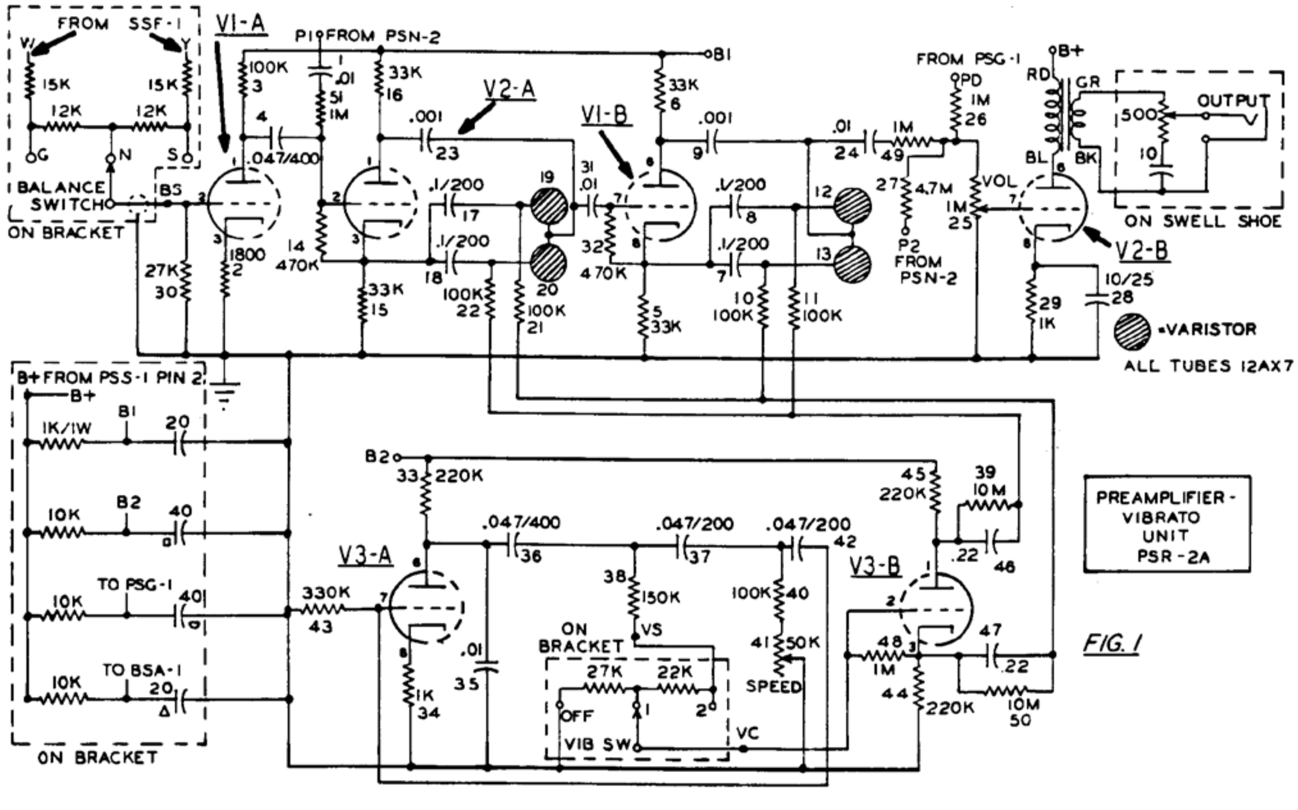


FIG. 1

With the PSR-2A circuit, the signal modulator stages V2A, V2B idle at only a few mA now. The varistor coupling caps are increased from 22nF to 100nF, which would extend low-frequency response. The coupling of the phase modulator to the next stage uses capacitive coupling and the driven tube has a local grid leak, which would help isolate any grid leakage current from the varistors. The LFO is simplified with no RC filtering on the way to V3B, and modified V3B loading and coupling to the modulator. The varistor 'biasing' scheme is simplified with the 10Meg bleeds from the LFO PI.



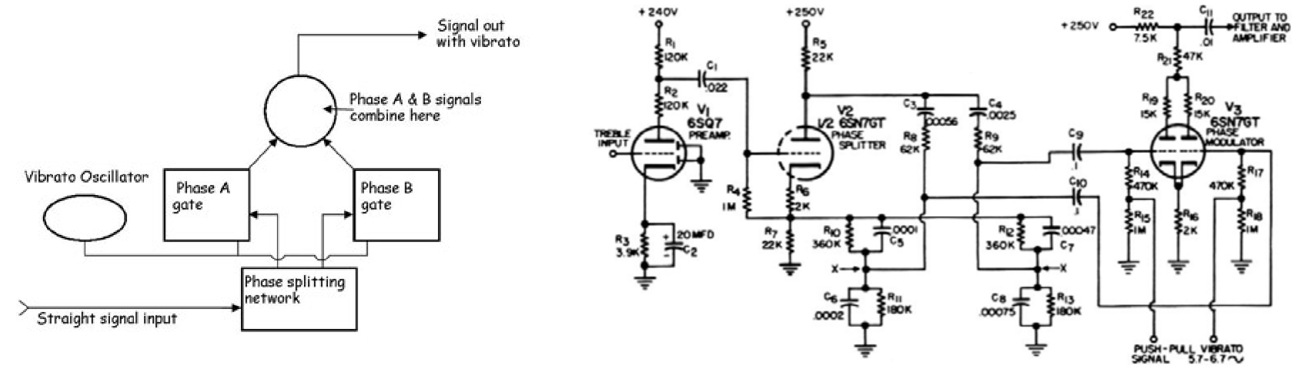
PREAMPLIFIER - VIBRATO UNIT PSR - 2A

FIG. 1

Appendix B: The Wurlitzer Vibrato Technique

It is worth briefly outlining the vibrato phase modulator technique used first in the Wurlitzer Model 44 organ [3] from 1953. The Wurlitzer technique predates Magnatone amps with Bonham's vibrato technique by about 4 years and has gone on to be used in many amps and effects right up to present times.

Dorf [4] presented the Wurlitzer vibrato circuit in the form of an effect 'pedal' in Radio & Television News April 1954 [8] for application to guitars and organs, and pointed to Moses' article for an explanation of phase shifting. Dorf also detailed the operation of the Wurlitzer organ in Radio Electronics, Jan 1955, which goes into detail about the vibrato circuitry.



[Courtesy of North Suburban Hammond Organ Society, Wurlitzer 4600 article, www.nshos.com/WUR12.htm]

[Graphic from 'Electronic Musical Instruments', by Dorf.]

Figure 19. Wurlitzer vibrato technique.

The cathodyne splitter valve V2 generates out of phase signals that are recombined by the circuitry R8-R13 and C3 to C8 to form 90 degree phase shifted signals. Those signals are then recombined by V3 with the signals being continuously amplitude modulated by out of phase LFO signals. The result is the original signal with an LFO frequency shift vibrato.

A few amplifiers, and many modern effects have since used the Wurlitzer vibrato circuit technique. For example, Jennings/JMI/Vox used it secretly from 1957 in their Vibravox, but most notably in the AC30 range (aside from the newer models that have dropped the channel) and also as a tremolo only circuit (see the VOX AC30 website article). The circuit technique spawned many interesting pedal names such as Magnavibe, Mindbender, TremO'Vibe, Tremster, Tube Wiggler, Vibromatic, Vibro-Stomp, VibraTrem, and Vibrotron.

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[3] wurlitzermodel44organservicemanual.pdf

[4] Dorf contributed a regular Audio Patents article to the Audio magazine from July 1950, and was identified as an Audio and TV Consultant from New York, and 'author of authoritative articles in leading radio publications' (prior to 1950). Dorf wasn't known to be involved in the Wurlitzer Model 44 vibrato circuit design.

SILICON CARBIDE VARISTORS

It has long been known that silicon carbide will, under suitable conditions of contact, exhibit a non-linear relationship between current and voltage. This may readily be demonstrated by measuring the voltage-current characteristic of a mass of small particles of silicon carbide compressed between metallic electrodes. As the voltage is increased from zero the current increases, at low voltage in direct proportion to the voltage and then much more rapidly. If the number of particles in the mass is large and the distance between electrodes large compared with the dimensions of the particles the non-linear resistance of the device is independent of polarity.

Experiments upon single particles with suitably made contacts indicate that the body resistance of the particle is small, ohmic, and independent of polarity.

The non-linear conduction exhibited by the mass of particles results from the voltage-dependent resistances at the point-to-point contacts between the granules of silicon carbide. The overall resistance characteristic may be thought of as due to large numbers of non-linear resistance contacts arranged at random in series and parallel. In a statistical sense the aggregate displays no dependence upon the direction of current flow. This varistor is an example of a "symmetrical non-linear resistor."

The simple device of containing a mass of silicon carbide particles under pressure between electrodes does not have the stability of characteristic under use conditions to afford wholly reliable circuit elements.

In 1930, McEachron (see *Journal A.I.E.E.*, Vol. 49, 410 [1930]) described a silicon carbide ceramic non-linear resistor to which the name Thyrite was given. The material consists of silicon carbide particles bonded in a ceramic matrix. Similar materials are known under various names such as Metrosil and Atmite.

The essential steps of manufacture are these: suitable silicon carbide particles, clay and water, sometimes with a minor constituent such as carbon, are mixed to form a plastic mass. The mass is partially dried and forced through screens to obtain a slightly damp granular powder. This material is compressed under high pressure into desired shapes, generally flat disks or rods. These pieces are further dried and heat treated in a reducing atmosphere at a temperature in the neighborhood of 1200 deg cent. The fired pieces are hard and strong and have mechanical properties quite similar to those of dry process porcelain. Electrodes on the opposite plane faces are provided by spraying or Schooping a layer of metal such as brass, copper, aluminum, or tin. The piece is then usually impregnated with a moisture-repellent organic substance to prevent pickup of water, which adversely affects their electrical stability.

The electrical properties of the product are profoundly affected by the parameters of process: materials, particle size, moisture content, forming pressure, and especially temperature, time, and atmosphere of the heat treatments. The products of different manufacturers differ somewhat in electrical properties, most importantly in the degree of non-linearity, and the characteristics of Fig. 8 are to be taken only as generally indicative. The current-voltage characteristic shown is closely represented by the equation

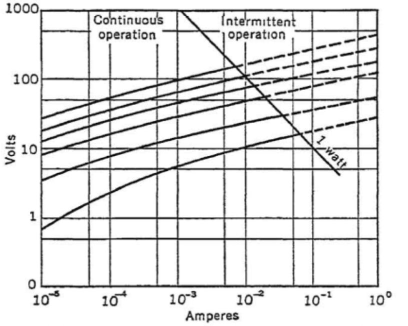

FIG. 8. Representative D-c Characteristics of Some 3/4-in.-diam. Silicon Carbide Varistor Disks

$$I = C_1E + C_2E^n$$

where I = current through the piece, E = voltage applied to the piece, C_1 and C_2 are constants depending on the material and geometry of the piece, and n is an exponent the value of which depends on various factors in the manufacturing process and generally lies between 3.5 and 5.0. Some manufacturers indicate values of n as high as 7.0 but only for pieces having resistances much above the range indicated in Fig. 8.

The variation of characteristic through control of manufacturing processes and geometry of the piece permits coverage of an enormous range of current and voltage. This range may be further extended by connection of pieces in series or parallel. It is to be noted from Fig. 8 that as the resistance of the piece decreases the value of n decreases also, and this being typical of all manufacturers' products may be considered an inherent characteristic of the presently made material. In consequence it is not possible with this device to obtain marked non-linearity at low voltage.

In common with semiconductors the silicon carbide varistor exhibits a negative temperature coefficient of resistance. The coefficient does not have a single value but varies both with the material and with voltage and temperature. The values of the coefficient at constant voltage cover a spread of from 0.3 per cent to 0.9 per cent per degree centigrade in the normally used range of temperature. The higher values of temperature coefficient are observed at the lower voltages.

At high frequencies consideration should be given to the presence of a capacitance effectively in parallel with the non-ohmic resistance. The exact value of this capacitance is determinable only by measurement, but the order of magnitude may be calculated by assuming the material to have a dielectric constant of 30 to 200.

Commonly used shapes are rods and disks. Small disks and rods may be furnished with leads soldered to the metallic electrodes on the faces of the piece. Disks are also made with holes in the center and clamped together with wiring terminals by means of a central bolt. Disks and rods of all sizes are used with spring clip mountings which furnish mechanical support and electrical connections.

When used under high humidity conditions, or at low currents, the organic impregnant, referred to in the description of the fabricating process, may not be sufficient protection against moisture and further precautions may be necessary.

Approximate values of mechanical and thermal properties of importance in circuit element design are as follows:

Bulk density	2.35 grams per cu cm
Compression strength	15,000 to 23,000 lb per sq in
Specific heat	0.17 to 0.21 cal per gram per deg cent
Thermal conductivity	0.0034 cal per cm per sec per deg cent

Requirements on the current-voltage characteristic for a particular application may be stated in a number of ways; the following are commonly used.

(*a*) The voltage E_1 at a current I_1 shall be greater than some value, and the voltage E_2 at a current I_2, where I_2 is greater than I_1, shall be less than some value. This statement of requirements contains implicitly a requirement as to the minimum value of n.

(*b*) The voltage at a given current I shall be equal to a value $E \pm X$ per cent, and the value of n shall lie within certain limits throughout a range of current.

It is to be noted that considerable differences in characteristic may exist between pieces meeting a set of such requirements. In commercial manufacture the range of voltage at a given current commonly runs ±20% about the average. Accuracy of meters used in checking requirements is important since errors in voltage readings are to be multiplied by n in determining their effect on current readings.

Self-heating resulting from power dissipation in the varistor lowers its resistance (negative temperature coefficient of resistance), but this effect is in general reversible; that is, no permanent effects on the characteristic are produced by moderate heating, say from 100 to 150 deg cent. The safe upper limit of heating is oftentimes determined by the moisture-resistant organic compound used as an impregnant. As shown in Fig. 8, 1.0 watt for a disk of 3/4-in. diameter suspended in free air at 50 deg cent is a limit recommended by one manufacturer. Very heavy transient currents may alter permanently the characteristic, usually in the direction of decreasing the resistance.

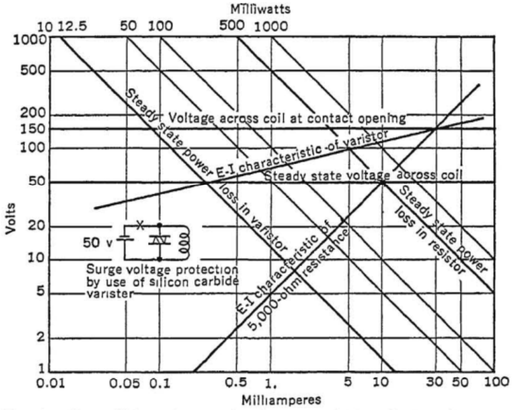

FIG. 9. Surge Voltage Protection by Use of Silicon Carbide Varistor

APPLICATIONS. (1) A silicon carbide varistor connected across the terminals of an electromagnetic winding acts to limit the surge voltage generated when the field is opened. As shown in Fig. 9 the maximum value of voltage across the varistor may be determined from the point on the voltage-current characteristic corresponding to the steady-state value of current I_0 in the winding. As compared with an ordinary resistance shunt across the winding to secure the same voltage-limiting effect, the varistor dissipates much less power when the coil is steadily energized.

(2) In certain carrier telephone system filters exposed to high incoming voltage, the condenser of a high-Q combination of coil and condenser has been protected by a varistor in shunt.

(3) Some of the smaller telephone switchboards have line lamps connected directly in the subscriber's loop for signaling. These line lamps are exposed to electrical disturbances that may be impressed on the outside lines, and if the disturbances are severe enough the lamps may be burned out. Silicon carbide varistors have been used very effectively in parallel with the lamp to bypass large incoming surges. The high resistance of the varistor at the normal signaling level has no appreciable effect on the lamp illumination.

(4) Use is made of varistors to protect contacts controlling inductive circuits from the deleterious effect of sparks resulting from the opening of such circuits. Usually the varistor is connected across the winding rather than across the contact to avoid continuous current drain. Though such an arrangement is useful it is not a satisfactory general solution of the problem. The varistor increases the release time of the relay or switch magnet, though not to the extent that an ohmic resistance of equivalent spark quenching action would do, and it does not entirely eliminate high-frequency oscillations across the opening contact due to the associated wiring.

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The introduction of variable resistors made from silicon carbide into the telephone set signaled the beginning of high-volume low-cost production of these devices by the Western Electric Company. Although these variable resistors, known as varistors in the Bell System, were introduced into telephone circuits in the early thirties, their use in a volume regulating circuit of the new telephone set was the first high-volume application in the Bell System. Today, two sizes of varistors used in this application account for more than 90 per cent of our current annual production of 15 million silicon carbide varistors. It is the purpose of this talk to describe the manufacturing process and facilities used to produce these units at a low cost.

The use of the electrical characteristics of silicon carbide presents three basic manufacturing problems. First is the effect of the abrasive qualities of silicon carbide upon the wearing surface of machines used for varistor manufacture. Second is the difficulty in determining the suitability of given lots of raw material for use in the manufacture of the varistor. The third difficulty encountered in the manufacture of the varistors is maintaining the proper balance between the various process parameters to insure maximum yield.

The slides used in connection with this discussion will include an outline of the manufacturing process and pictures of some of the facilities used. Such interesting process features as rubber-faced punches for pressing disks, automatic metalizing for application of contacts, vacuum pickup to minimize wear on an automatic test set, and paper-clip type of terminals to facilitate soldering will be discussed.