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# Optimizing Output Filters Using Multilayer Polymer Capacitors in High Power Density Low Voltage Converters

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## ABSTRACT

At lower output voltages, ripple filter capacitors consume larger portion of converter volume as switching frequencies rise to achieve higher power densities. The impedance/volume of Multi-layer Polymer (MLP) capacitors is now comparable with the best of other technologies above about IMHz at a lower cost, and the advantage will increase as capacitors with dielectric films thinner than one micron enter production.

The principal problem with using MLP capacitors to reduce filter volume is low energy storage. Volumetric capacity is presently similar to that of X7R Monolithic Ceramic (MLC) capacitors, which is about an order of magnitude less than that of aluminum electrolytics, and nearly two orders of magnitude less than solid tantalums. The technical challenge to exploit the high frequency ripple attenuation capability of MLP capacitors while overcoming their limited "energy reservoir" capability. The successful application of MLP caps to low voltage filters was found to require a suitable combination of converter, control and filter design.

## INTRODUCTION

Two major trends in Switchmode converters are the lowering of logic supply voltages below 5V, and higher power densities through higher conversion frequencies. These trends meet and clash in low voltage DC-DC converters in distributed power systems. The field has been rife with new technological problems and solutions; low voltage schottkys and synchronous rectifiers, lower loss ferrites, a better understanding of eddy current losses in windings, and new converter topologies to name a few.

A problem less well addressed has been low voltage output filters. Scaling a filter design to lower voltages at the same power level requires that filter element impedances vary as the inverse square of the output voltage. Low value/high current inductors are little problem, but very low impedance capacitors are another matter.

Theoretically capacity should vary as the inverse square of voltage as dielectric thickness is reduced, but the higher frequency impedance (per unit volume) of virtually all capacitor types tends to "bottom out" somewhere in the 20V to 50V range. The limitation is ESR (Equivalent Series Resistance) in electrolytic capacitors; although capacity increases at lower voltages, ESR remains relatively constant. In plastic (MLP) and ceramic (MLC) caps, capacity is technologically limited by dielectric thickness and dielectric constant.

Below this characteristic "bottomed out" voltage filter capacitor volume grows as the inverse square of voltage, unless either better capacitors are used or the filter is redesigned to require less "capacity". A point is quickly reached where significantly better capacitors are simply not available. It then becomes necessary to choose the best capacitor technology for the application, and try to minimize the capacity required.

## ALTERNATIVE CAPACITOR TECHNOLOGIES

With their large specific capacities, electrolytic caps have traditionally been used for ripple filtering. Aluminum electrolytics were preferred in cost sensitive commercial applications, with the more expensive tantalums used in high temperature or high performance military/aerospace environments.

Aluminum electrolytics in particular have seen tremendous improvements in performance (shelf life, ESR, energy density, leakage, lifetime, and reliabilty) in recent decades. Nonetheless, ESR still typically limits filter performance above a few KHz. [Although in some of the highest technology devices, such as the Sanyo OS-Con series, ESR is sufficiently low that its effects are held off until ESL (Equivalent Series Inductance) becomes the dominant impedance above a few hundred KHz.]

Solid tantalum capacitors have much higher specific capacities and lower ESRs, although at a premium price.(Wet tantalums require a hermetic case and seal, which increases HF impedance prohibitively.) Some surface mount designs (e.g. the Sprague 195D series) also have very low ESL. Unfortunately, solid tantalums have been historically prone to failing short under high pulse currents, which can easily occur in filter applications. Despite this potential reliability problem, solid tantalums have found significant application in lower power (tens of watts) converters at 100KHz to 1MHz conversion frequencies.

Surface mounted MLC capacitors have found some use at frequencies ≥ 1MHz, where their impedance can be less than that of electrolytics. Price has been something of a limiting factor in commercial applications, but the major technological problem has been the construction of semi-stable capacitors larger than luF which can be surface mounted on printed circuit boards. Even if they survive the stress of soldering, temperature cycles tend to crack larger ceramic capacitors due to coefficient of thermal expansion mismatch between capacitor and substrate.

Metallized plastic capacitors have similar electrical properties to MLCs without the differential expansion problem, but until recently they have been rather larger then MLCs of the same capacity. This has changed with thinner dielectrics and new construction techniques; they are now volumetrically competitive with ceramic capacitors and will become more so in the future.

**Comparing Capacitor Technologies**

It is difficult to make meaningful comparisons from published data, so in order to make an "apples to apples" comparison I have taken my own data on the best devices (to my knowledge) of each type. Impedance vs frequency was measured for capacitors of essentially the same volume with as similar as possible (low inductance) mounting techniques.

Four terminal transimpedance (Vout/Iin) was measured on parallel copper traces on a 0.062" thick board, with the "Lo" side connected to a ground plane on the far side (details available on request). More or less conventional surface mounting was used with all capacitors except the aluminum electrolytic, which was measured first with leads coming through the ground plane to the traces (0.06" leads), and then with copper foil tabs soldered at the case and "surface mounted" for the lowest possible ESL. The results of the measurements are shown in Figure 1, with further details of the capacitors and mounting tabulated in Table 1.

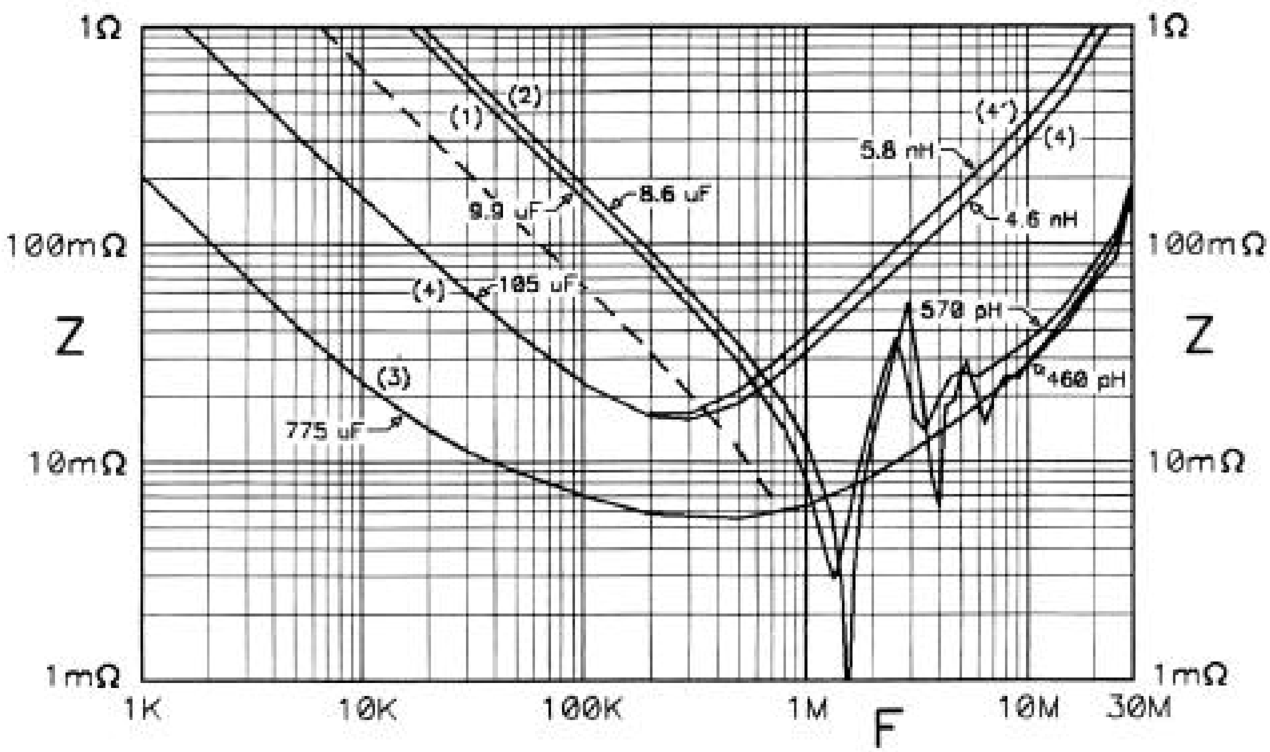



**Fig. 1. Capacitor Impedance vs Frequency**

Table 1

| Test # | Cap. Type | Cap. Qty. | Rated: C: | Rated: V: | Manufacturer | Part # | Total Measured Capacity: |
|---|---|---|---|---|---|---|---|
| (1) | MLP | 1 | 10uF | 50V | ITW Paktron | 106K050C5S4 | 9.9uF |
| (2) | MLC | 10 | 1uF | 50V | (?) | (?) | 8.6uF |
| (3) | Tantalum | 15 | 47uF | 10V | Sprague | 195D476X0010Z2T | 775uF |
| (4) | Aluminum | 1 | 100uF | 10V | Sanyo OS-Con | 10SA100K | 105uF |
| (4') | " | " | " | " | " | " | " |

| Test # | Cap/Array Dimen. L: | W: | H: | Volume: Cu. In. | cc: | Comments: | EstimatedCost |
|---|---|---|---|---|---|---|---|
| (1) | 0.05" | 0.40" | 0.22" | 0.044 | 0.72 | Dimen. w/o epoxy | $4.0 |
| (2) | 0.50" | 0.23" | 0.375" | 0.043 | 0.71 | 2stacks of 5 caps | $6.0 |
| (3) | 0.52" | 0.28" | 0.31" | 0.045 | 0.74 | 5W x 3H cordwooded | $7.5 |
| (4) | 0.32" | D x 0.042"L | | 0.043* | 0.71 | 0.00" lead length | $0.6 |
| (4') | " | " | " | " | " | 0.06" lead length | " |

* Rectangular volume

It can be seen from Figure I that the impedance of the MLP and MLC capacitors tested is virtually identical, with most of the difference attributable to the MLC's capacity being near the low end of their ±20% tolerance. (The projected impedance of future MLP caps using 0.6 micron film is shown with the dashed line.) Above 1MHz the impedance is also very similar to that of the solid tantalum capacitors, except for the transmission line resonance effects described and explained in an earlier paper [1]. These resonances can be removed by orienting the capacitor plates normal (rather than parallel) to the mounting surface.

Other comparisons between the technologies are:

**1) Aluminum Electrolytics:**

These have the lowest cost per uF of all the capacitors considered. Although ESR is higher for the same volume, they are competitive when size is less important than cost. The density of aluminum electrolytics is significantly less than that of the other types, but this is largely offset by the higher impedance as we shall see below.

The ESL of the leaded construction limits viability at frequencies above a few hundred KHz. I have also received unconfirmed reports of catastrophic failures in aluminum electrolytic capacitors with organic semiconductor electrolytes, possibly caused by high temperatures and/or reverse voltage.

**2) Solid Tantalum Electrolytics:**

Solid tantalums have the highest specific capacity, but also the highest price for a given impedance above 1MHz. Maximum operating temperature is high: 125C with voltage derating.

Solid tantalums tend to fail short circuit with overvoltage or high pulse currents, which can be catastrophic if sufficient unfused fault current is available. Physical size is limited, so large banks of capacitors are required at higher power levels, which has a negative impact on reliability.

**3) MLC Capacitors:**

Ceramic capacitors generally have the highest temperature capability, and may have the lowest ESR (depending on the construction and ceramic type). Larger capacitors (>1uF with X7R dielectric) are prone to cracking and catastrophic failure under thermal cycling with mismatched thermal coefficient of expansion substrates. Arrays of smaller capacitors with lead frames are being used to overcome this problem.

The palladium silver electrodes used at low voltages make them intrinsically expensive, with the price dependent on raw materials costs. Efforts to raise specific capacity with thinner dielectrics raises dissipation factor (and hence ESR), and problems have been experienced with silver migration and short circuiting.

**4) MLP "Plastic" Capacitors:**

These are the lowest cost capacitors of the low HF impedance capacitor types. No expensive raw materials are used, so costs can be expected to hold or decline with improvements in film manufacturing technology.

The self healing capability of metallized plastic capacitors make them quite reliable, with relatively graceful and typically gradual failure with overvoltage. [I once successfully (and somewhat dramatically) demonstrated the robustness of this self healing property to a skeptical fellow engineer by cutting the corner off of a capacitor with a hacksaw, and then driving a nail through a drilled hole in the middle. About 25% of the capacity was lost, but it would still withstand rated voltage.]

ESR is very low, typically 3 milliohms at 1MHz for a 10uF, 50V capacitor, and can be lowered further if needed with heavier metallization (which will have minimal effect on capacitor volume). RMS current capability is correspondingly high, on the order of 1A/uF. Voltage and current ratings apply to 85C, where ESR is at a broad minimum with polyester dielectric; operating temperatures to 125C are possible with derating.

MLPs are also the least massive or dense of the low HF impedance capacitor types, with a density about half that of the solid tantalums and nearly 1/3 that of MLC caps, which is an advantage in portable or mobile applications. These mass and mass- impedance product vs frequency of the capacitors tested is given in Table 2 below. A low mass-impedance product means that a lower capacitor mass is required to achieve a given impedance.

## FILTER DESIGN USING MLP CAPACITORS

We have seen that MLP capacitors have very low impedance at higher frequencies, which makes them attractive for high frequency filtering with minimum volume. They cannot be use to directly replace electrolytic capacitors, however, nor can one design filters in the same way.

### The Problems Involved

Electrolytics for filtering are principally selected to tolerate the ripple currents involved. (As the saying goes, you buy the ESR, not the uF.) Ripple voltage is then usually not too far off the target, and as long as a reasonably high feedback gain bandwidth (F.B. GBW) or unity gain crossover frequency is achieved, the bulk capacity is normally sufficiently high that transient load response is not a problem. On the other hand, if one buys just enough MLP capacitors to handle the ripple current, it initially looks like very little capacity is needed, but several problems soon become evident.

First, the ripple voltage is somewhat higher than with electrolytics of the same ripple current capability. The relatively high ESR of electrolytics limits ripple current, and thus ripple voltage. The low ESR of MLP caps allows high ripple currents with much lower capacity, so the higher (reactive) impedance results in higher ripple voltage.

Second, any attempt to reduce ripple voltage by using a larger inductor instead of more capacity gives the output filter a high surge impedance. When the energy stored in the inductor at full load becomes comparable to the energy stored in the filter capacitors, no control circuit can prevent large transient output voltages with sudden changes in load current.

Third, even if the capacitors are made large enough to limit ripple voltage with a reasonably small input choke, the total capacity will still be one or two orders of magnitude less than with typical electrolytics, with correspondingly low “reservoir” energy storage. Sudden large changes in load current will cause the output voltage to rise or fall rapidly, and even the fastest conventional voltage feedback loop will be too slow to prevent significant transient output voltages.

Multi-stage filters can reduce ripple without increasing total L or C, but this creates problems in the voltage feedback loop. The additional L-C filter stages will have a high Q with low loss MLP capacitors, which will cause the feedback loop to oscillate if the loop gain rises above unity due to resonant peaking. Lowering loop gain to achieve stability will also slow down the transient response to load changes, which is already a problem.

Damping can be used to reduce resonant Q and allow higher feedback gain, but too much damping will cause excessive phase shifts at frequencies well below resonance and again require lower feedback gain for stability [2]. There are also problems with achieving this damping without significantly increasing the capacity required, losing filter ripple attenuation and/or causing excessive filter damper losses.

Research to find economical and viable low voltage filters using MLP capacitors was funded by ITW Paktron. SPICE modeling proved invaluable in this exploratory phase, with about 150 filter configurations and variations investigated. Representative results and the most promising filters found are presented in this paper.

### The Design Example Researched

As a design engineer I found it conceptually difficult to deal with an abstract “general case” or normalized filter. Since a proof-of-concept prototype using the research result is planned for exhibition at the HFPC ‘95 conference, I used the target requirements for that converter and filter as the research example. Viable filters for this specific application could then be transformed to other operating frequencies, voltages and currents using the simple formulas provided later.

The intent is to develop filter circuits for low voltages, so a converter output of 3.3 VDC and 60A (200W) was selected. A 1MHz conversion frequency is representative of leading edge practice, and an output ripple target of I% P-P (33 mVp-p) was suitably challenging. After initial feasibility studies I decided to generally limit total HF capacity to 60uF (1uF/A out), and the main filter inductance to 30 nH (2 x Icrit

Table 2

| Test # | Cap. Type | Volume cc: | Mass: | S.G. g/cc: | Mass-impedance (g-millohms) @: 10KHz | 100KHz | 1MHz | 10MHz |
|---|---|---|---|---|---|---|---|---|
| (1) | MLP | 0.72 | 1.71g | 2.4 | 2740 | 274 | 14 | 62 |
| (2) | MLC | 0.71 | 4.24g | 6.0 | 7840 | 784 | 53 | 122 |
| (3) | Tantalum | 0.74 | 3.47g | 4.7 | 80 | 24 | 22 | 97 |
| (4) | Aluminum | 0.71 | 1.06g | 1.5 | 174 | 24 | 34 | 314 |

operation at full load with D = 0.5). In summary, the researched filters were designed for:

1) A conversion frequency of 1MHz;
2) 3.3V, 60A output;
3) 60uF total filter capacity;
4) 30nH main filter inductor;
5) Output ripple of 1% (33mV) p-p at D = 0.5;
6) Load transient response to be evaluated.

**Converter and Control Requirements**

The energy stored in the 60uF output capacitors at 3.3 V is 327 microjoules, which is even theoretically only enough to supply full output power for 1.65 us! With a switching frequency of "only" 1MHz the design task may look impossible, but it is not. However, the difficulties of meeting output ripple, feedback stability and load transient response goals place strong constraints and requirements on the converter, control and filter design. Extremely fast response to load current changes is paramount, which requires that:

1) The converter must be buck derived, in order to avoid the delays of the right hand plane zero in the response of boost and buck-boost topologies;
2) Current mode control must be used to remove the pole of the first output filter inductor from the feedback path;
3) The gain and phase characteristics of the filter must be optimized to allow for:
4) The highest possible gain bandwidth (GBW) in the voltage feedback loop (the feedback GBW asymptote determines settling time), and even then:
5) Output current feedforward (which senses changes in load current and adds the information to the current reference) is mandatory for acceptable transient load response;
6) The filter input inductor must be kept small to limit inductive energy storage relative to that in the capacitors, and to achieve a high current slew rate.

**Feedback Loop Stability Criteria**

The conventional gain and phase margin criteria are inadequate when evaluating feedback stability around multi-stage filters; depending on degree and type of resonant damping, gain-phase relationships can range from benign to pathological. The Nichols chart is the tool of choice here, which plots loop gain vs phase with contours of stability.

We can't afford to "waste" much feedback gain with such low energy storage, and as filter and other circuit values are fairly well controlled (5% to 10% tolerance), I chose the equivalent of a 45 degree phase margin as my stability criteria. This is essentially the G/(1+G) = 2 dB contour on the Nichols chart, which crosses the gain axis at 5.1 dB "gain margin". This is somewhat less than many designers would feel comfortable with, but in practice a gain-phase trajectory tangent to the 2 dB contour typically crosses the gain axis somewhat lower down (as well as crossing the phase axis with more than 45 degrees phase margin).

## FILTER CIRCUIT DISCUSSIONS AND EVALUATIONS

The filter circuits discussed are shown in Figures 2 to 15. Although not shown in the figures, all HF filter capacitors were modeled with an ESR equivalent to 1 milliohm in 30uF. Key performance parameters are listed in Table 3, and filter "surge" voltages are plotted in some detail for selected filters in Figures 16 and 17. A filter rating system is applied, with one "star" for filters meeting minimum target requirements through three stars for the best filters found in the study.

Table 3

| Filter | | | Ripple | | Surge | Max. | Filter Losses, W | | | |
|---|---|---|---|---|---|---|---|---|---|---|
| Fig. # | Leg. Dmp. | Notch Comp. | mVp-p 0.5 | D= 0.25 | Volt 1/1us | F.B. GBW, | Cap ESR | Dmpr Res. | Ntch Res. | Total |
| 2 | - | - | 33 | 47 | 1.31 | 245K | 0.02 | - | - | 0.02 |
| 3 | Ind | - | 65 | 92 | 1.05 | 99K | 0.91 | 0.30 | - | 1.21 |
| 4 | Ind | - | 61 | 86 | 0.98 | 111K | 0.90 | 0.22 | - | 1.12 |
| 5 | Ind | L | 32 | 46 | 0.86 | 138K | 0.83 | 0.63 | - | 1.46 |
| 6 | Ind | L - R | 2.2 | 12 | 0.91 | 130K | 0.86 | 0.16 | 9.90 | 10.9 |
| 7 | Ind | L | 1.7 | 9 | 1.03 | 100K | 0.95 | 1.22 | - | 2.17 |
| 8 | Ind | L | 1.8 | 9 | 1.18 | 94K | 0.95 | 1.22 | - | 2.17 |
| 9 | Cap | - | 81 | 114 | 0.97 | 64K | 1.16 | 1.01 | - | 2.17 |
| 10 | Cap | L | 29 | 44 | 0.86 | 99K | 1.07 | 0.94 | - | 2.01 |
| 11 | Cap | - | 86 | 121 | 0.60 | 91K | 0.81 | 0.91 | - | 1.72 |
| 12 | Cap | - | 31 | 44 | 0.29 | 79K | 0.27 | 1.39 | - | 1.66 |
| 13 | Cap | - | 97 | 137 | 0.43 | 27K | 0.90 | 1.31 | - | 2.21 |
| 14 | Cap | L | 62 | 90 | 0,39 | 37K | 0.74 | 1.08 | - | 1.82 |
| 15 | Cap | L - R | 4.1 | 20 | 0.40 | 35K | 0.84 | 1.22 | 18.5 | 20.6 |

**Filter Performance Parameter Definitions**

The “P-P ripple” voltage has been calculated as the RMS of the P-P fundamental, 2nd and 3rd harmonics, for both a “worst case” duty cycle of 25% (high input voltage), and a “typical” duty cycle of 50%; effects of higher harmonics are negligible. True P-P ripple would depend on relative harmonic phase angles.

“Surge voltage” is the modeled peak output voltage rise due to a FL to NL change in load current over a finite fall time “Tf”. Assuming a current mode controlled buck converter with output current feedforward, the converter response to the load change will be a drop of the filter input voltage to zero after a time delay “Td” ranging from zero to one switching cycle, depending on where in the switching cycle the current fall starts. The tabulated surge voltage is for a 100%-0% current fall time Tf = lus, and voltage fall delay time Td = lus.

The “Maximum Feedback Gain Bandwidth” (F.B. GBW) is essentially equivalent to the feedback unity gain crossover frequency as a measure of the voltage feedback loop response speed and settling time. For these filters it was calculated as the maximum 10KHz loop gain times 10KHz, with the gain-phase curve on the Nichols chart tangent to the G/(1+G) = 2dB contour.

“Filter Losses” were calculated AC losses for the ripple fundamental at D = 0.25 for filter capacitors (assuming one milliohm ESR for a 30uF cap), and for any intentional damper or notch inductor phase shifting resistances. Calculated losses do not include filter inductor or circuit conductor Iosses.

**Filter Circuit Discussions**

Figure 2 is a single stage filter used for performance comparisons with the remaining multi-stage filters. Filter capacity is the same 60uF of total HF capacitance used in all the filters studied, with an input inductor large enough (120nH) to limit output ripple to 1% P-P at D = 0.5. Performance is acceptable, although the F.B. GBW needed is 1/4 the conversion frequency. Filter capacitor ESR losses are negligible.

Figures 3 & 4 are similar inductive leg damped filters, using a tuned L-R-C and untuned L-R dampers respectively. Their relative performances (Table 3) illustrate my general observation that (contrary to my expectations) a tuned damper has no advantage over an untuned damper, at least when used in the inductive leg. (Potential benefits of tuned damping in the capacitive leg have not been investigated.) These filters have a lower surge voltage than Figure 2, but ripple voltage is about twice as high.

Figures 5 & 6 add an inductor “Ln” from the input to output of the filter of Figure 4, which produces a dip or notch in the filter response at the fundamental conversion frequency of 1MHz through cancellation of the currents through the two paths to the output. The effect in the Figure 5 filter is to reduce the ripple to the target level of 1% P-P, while further reducing surge voltage and improving F.B. GBW slightly over the Figure 4 filter. The addition of a phase shifting resistor “Rn” in Figure 6 produces a deep notch and reduces ripple considerably, but at the expense of about a 5% of full load power loss in Rn.

The filter of Figure 7 uses an additional filter stage to increase the phase shift in the power path, allowing a deep response notch at the switching fundamental without the need for the lossy phase shifting resistor Rn. Ripple is even lower than the Figure 6 filter, although surge voltage and GBW are now back near those of Figure 4. Although the filter of Figure 7 is locally “optimal”, the “Rd” damper resistors have unrealistically low ESL requirements (<<1nH). In Figure 8 the filter is redesigned to allow inductance in series with Rd, although with some further compromise in surge voltage and GBW.

A representative capacitive leg damped filter is shown in Figure 9, where the total capacity is held to 60uF. The performance is generally worse than in the roughly equivalent inductive leg damped filter of Figure 4: surge voltage is similar while ripple is higher and F.B. GBW lower. (Increasing the size of the damper capacitors and the second stage inductor L2 reduces ripple moderately, but with a lower F.B. GBW and a rapidly increasing surge voltage).

A response notching inductor “Ln” is added to the Figure 9 filter in Figure 10, which improves all performance parameters considerably. Surge voltage and ripple are now comparable to the filter of Figure 5, with a moderate shortfall in F.B. GBW.

The “obvious” solution to capacitive leg damping is to use relatively small and cheap electrolytic capacitors for damping instead of the expensive low ESR capacitors. This alternative was explored by assuming electrolytics were sufficiently “free” or at least cheap in terms of cost and volume that they could be simply added to the 60uF of HF capacity used in this study.

Out of a total of twelve “electrolytically” damped filters studied (with Cd ranging from 10uF to 480uF), two representative filters are shown. Relatively small damper caps (15uF) are used in Figure 11, with rather larger caps (120uf) in Figure 12; damper resistances Rd are optimized for maximum F.B. GBW in each case. Ripple in the Figure 11 filter is too high, while that in Figure 12 meets the target. Surge voltages are also considerably improved over the other filters considered, although F.B. GW are a little low (i.e, longer settling time).

The situation is not quite what it seems, however; this is one of the places where the decision to use a “real world” filter requirement bore fruit. The 5.4 milliohm optimum damper resistance in Figure 12 (= max Cd ESR) is not remotely available in any suitable 120uF electrolytic that I could find. Thus the “questioned star” rating of Figure 12; the filter would work if it were realizable.

A close alternative was to use four Sprague 195D series 47uF, 10V tantalum capacitors in parallel for each damper, as shown in Figure 13. The typical measured ESR is 80 mil-

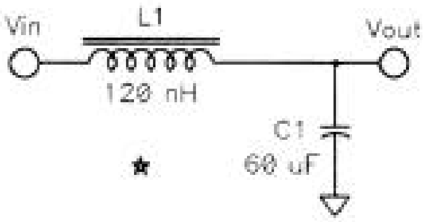



**Fig. 2**

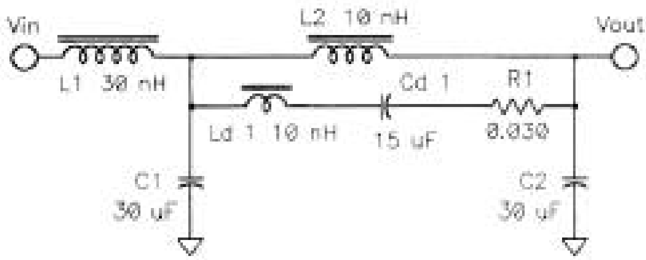



**Fig. 3**

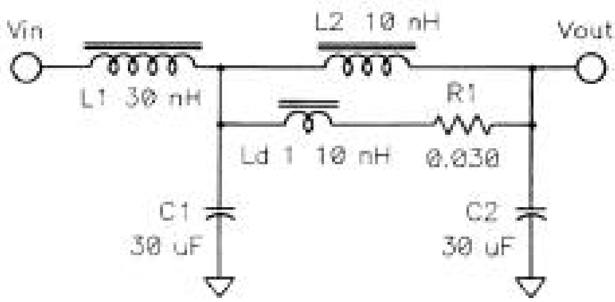



**Fig. 4**

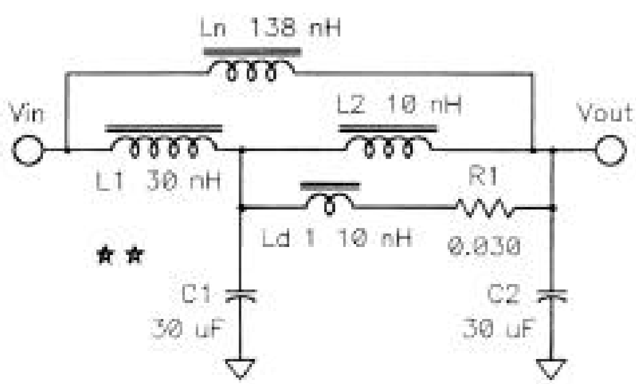



**Fig. 5**

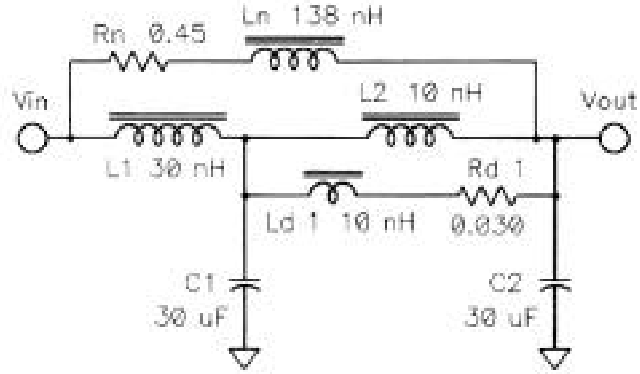



**Fig. 6**

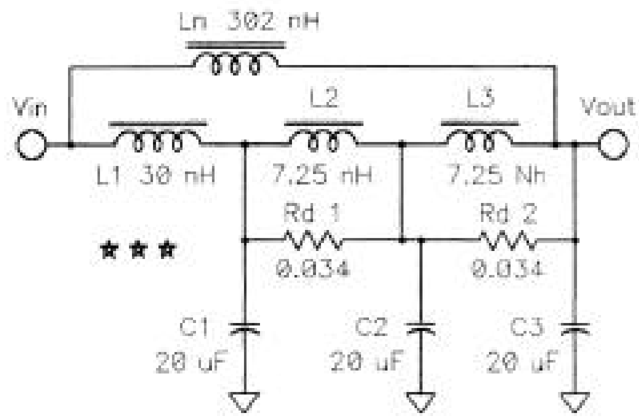



**Fig. 7**

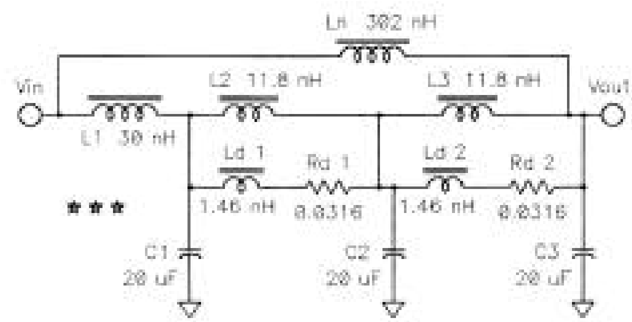



**Fig. 8**

**Figures 2 - 8. Single Stage, and Inductive Leg Damped Multi-Stage Filters**

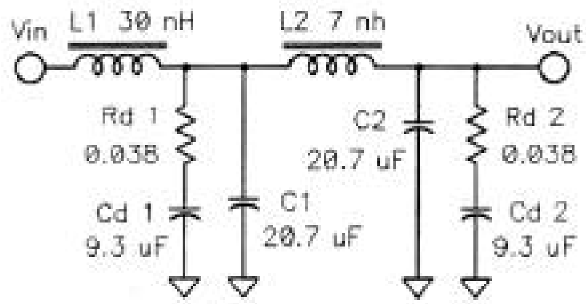



**Fig. 9**

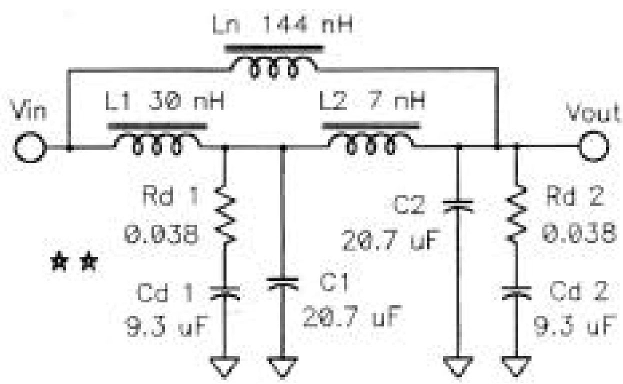



**Fig. 10**

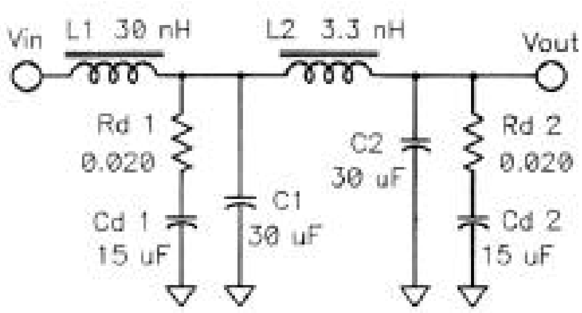



**Fig. 11**

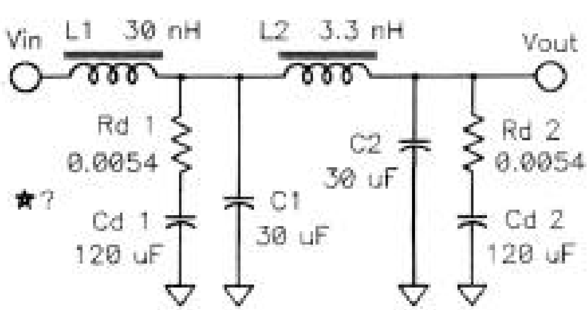



**Fig. 12**

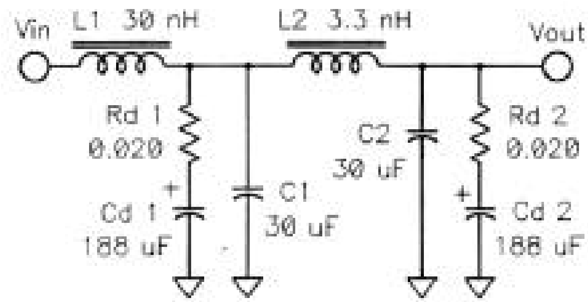



**Fig. 13**

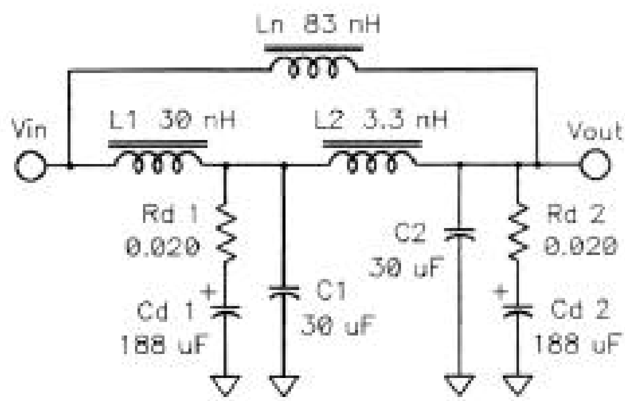



**Fig. 14**

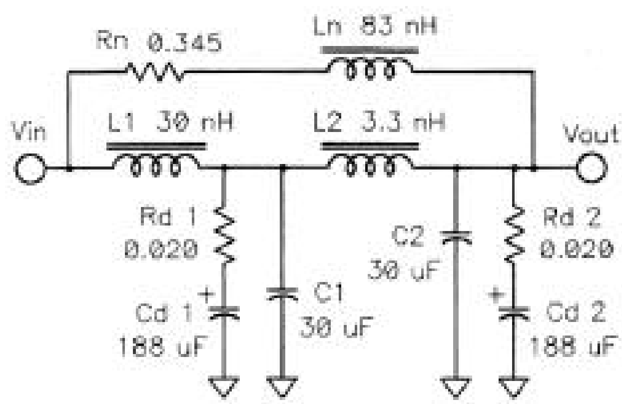



**Fig. 15**

**Figures 9-15. Capacitive Leg Damped Multi-Stage Filters**

liohms per capacitor, or 20 milliohms net for the 188uF bank of four, the same Rd as in the Figure 11 filter with Cd = 15 uF. The surge voltage is still quite good, but the F.B. GBW is reduced by a factor of 3 or 4 due to over-damping. The real surprise is that the ripple voltage is actually about 1dB higher than even the filter of Figure 11, which has all the same component values except 8 times smaller damper caps!

This counter-intuitive result of more ripple voltage with larger damper caps is principally due to the larger phase angle between the damper and filter capacitor currents. Although the damper current is about 40% higher with the larger damper caps, the phase angle is also about 25 degrees more resistive; the vector sum with the filter capacitor current is actually 6% less, and thus the combined impedance is about 6% higher. This occurs in each of the two filter stages, so total ripple is increased about 12% (pointing out the hazards of intuition, and the advisability of modeling the response of prospective filter circuits).

Even with the addition of a “notching” inductor in Figure 14 the ripple is still twice the target value; only the addition of a notch phase shifting resistor “Rn” brings the ripple voltage down below target, but the added dissipation in Rn is nearly 10% of the FL output power. Damping low impedance filters with electrolytics does not appear to be the way to go.

**Transient Load Surge Voltage**

Typically a “step” change in load current is used to evaluate transient response, but this is unrealistic for the voltages, currents and time frame involved with the filters studied. On the one hand it can be seen from Figure 16 that the output surge voltage improves dramatically as load current fall time increases from one to several microseconds.

On the other hand a 60A load current fall time of even 2us would cause a voltage surge of 1.5 V with a distribution bus inductance of only 50nH, even if the supply voltage were rock solid. Obviously heavy bypassing at the load is required when the current can change suddenly at low supply voltages, which slows down the rate of change of load current at the converter.

Characteristic improvements in surge voltage with shorter response time delays are shown in Figure 17 for the filter of Figure 7. Near instantaneous feedforward response reduces surge voltage, but is technically very difficult to achieve, and is only relevant with full load current changes between about 0.5 and 5us. For rational L.V. systems design, it will be necessary to define a realistic maximum converter output current dI/dt, the allowed maximum output voltage deviation, and the minimum input voltage for transient load increase measurements.

**Observations**

In the design of low voltage filters using MLP capacitors, my conclusions (which are guidelines and not gospel) are that:

1) A single stage filter meets the target requirements;
2) Overall performance can be improved with suitably damped multi-stage filters;
3) Electrolytic damper capacitors are not preferable to using

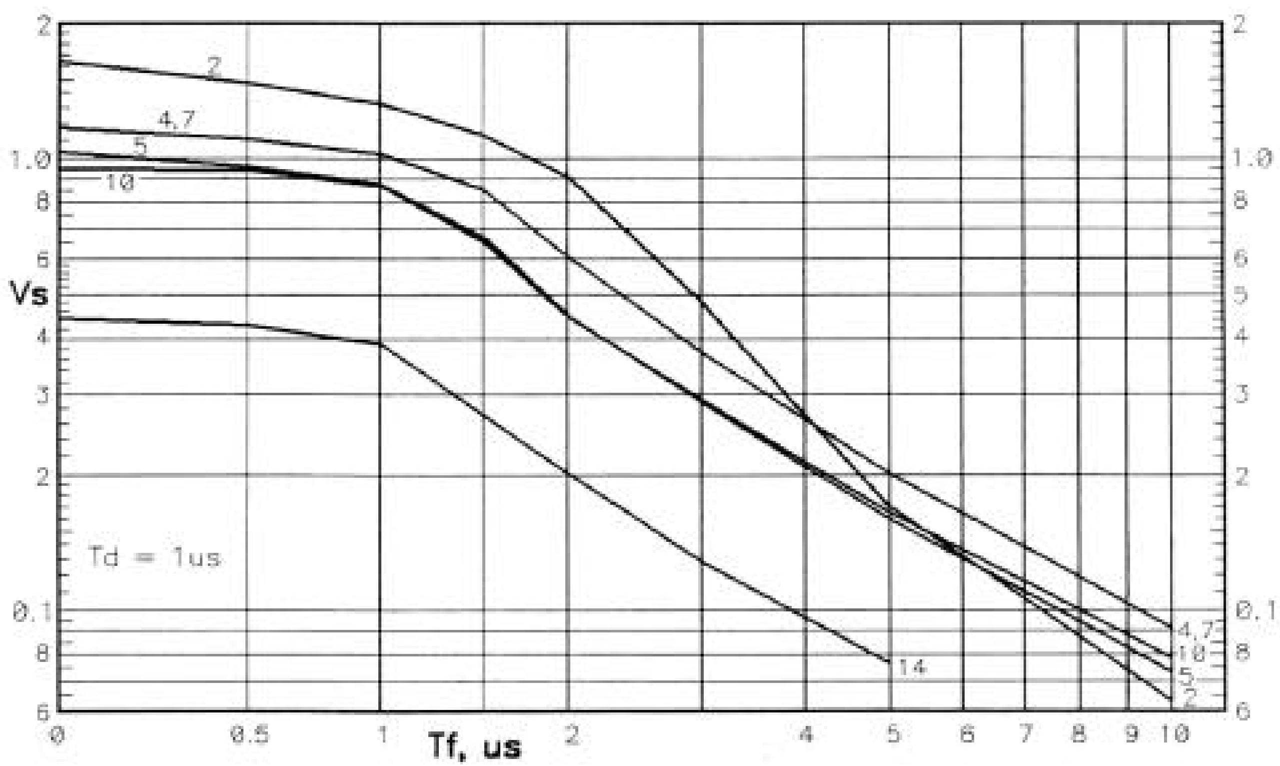



**Fig. 16 Filter Surge Voltage vs Current Fall Time “Tf” & Filter No.**

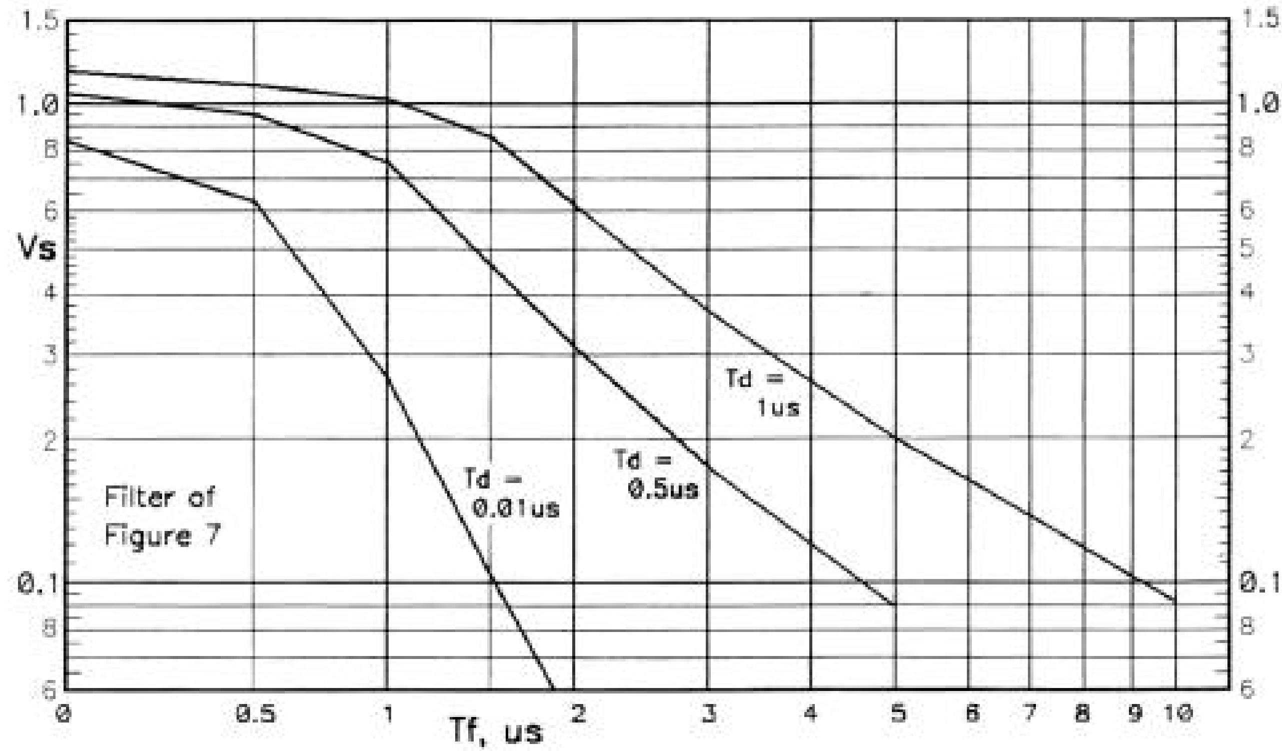



**Fig. 17 Filter Surge Voltage vs Current Fall Time "Tf" & Response Delay "Td"**

part of the allocated HF capacity for damping;

4) Inductive leg damping is preferable to capacitive leg damping;
5) Tuned L-R-C inductive leg dampers have no advantages over untuned L-R dampers;
6) An input to output inductor on a multi-stage filter can be used to create a response notch at the switching frequency, significantly lowering ripple and improving surge voltage and F.B. GBW (note that a larger "notching" inductor will raise the notch frequency);
7) Using a phase shifting resistor with a notching inductor to improve notch depth dissipates excessive power;
8) A constant conversion frequency is more or less essential, allowing the use of notch filtering and avoiding variations in transit delay in the switching cell.

**Inductive and Capacitive Leg Damping**

In an earlier paper [2] I had expected a duality between damping the inductive leg (with an L-R damper) and the more conventional capacitive leg damping (with a R-C damper). I have found in this research that this duality does not occur; inductive and capacitive leg damped filters behave differently, and the two types must be "optimized" individually and then compared. The reason seems to be that a current mode controlled converter an two stage filter respond as a C-L-C "pi" filter (with an AC current input and an essentially open circuited voltage output), losing the symmetry or duality between inductors and capacitors. The anticipated advantage of inductive vs capacitive leg damping was observed, however.

I found that my earlier published cures (Figure 4 in [21) for an "Optimum R-C Damper for Parallel L-C Resonant Circuit" could indeed be applied to (at least) two stage filters in order to calculate the optimum damper resistance for given filter and damper capacitors. The resonance to be damped is between the second stage inductor L2 and the two filter capacitors C1 and C2 in series, so the effective resonant inductance with one filter and damper capacitor is half the second stage inductance.

Due to the lack of duality noted, these curves do not appear adaptable for calculating the optimum inductive leg L-R damper resistance.

**Filter Construction and Layout**

Low voltage/high frequency filters have circuit impedances in the milliohm range at hundreds of KHz to MHz, and layout and construction will be crucial to performance. ESL of capacitors and damper resistors will be of particular importance. Although design guidelines are beyond the scope of this already overly long paper, a few words on damper resistors are in order.

As noted, the required damper resistances are in the milliohm range, and they may have to be fairly non inductive at high frequencies. Inductive leg damping has additional merits in this regard. First, the resistance in the parallel leg dampers used in figures 3-6 have some series inductance anyway, so the use of resistive sheet material might be practical. Alternatively, a higher valued resistor (≥1 ohm) could be "transformer coupled" across the inductor with a secondary winding, and the leakage inductance form part or all of the inductance in series with the damper resistor.

**Filter Transformations**

The filters presented can be transformed for operation at other currents, voltages and frequencies. If we define the output impedance as Z = V out/I out, then to transform a filter from Frequency F to F', and from Impedance Z to Z':

Let: Kf = F/F' and: Kz = Z/Z'

Then L' = L (Kz/Kf)  C' = C/(KzKf)  R'=R(Kz)

For each L, C, and R in the filter.

**Parallel to Series Inductive Leg Damping Transformation**

The parallel inductive leg dampers used for evaluation can be transformed to series inductive leg dampers as shown in Figure 18. Unlike a parallel to series L-R transformation, the parallel to series damper transformation is frequency independent.

**Other Variations**

Filter capacity is shown evenly divided between the stages of a multi-stage filter. Although modeling shows this to be "optimum", a 2:1 imbalance (1/3 and 2/3 split) in either direction (in the several two stage filters checked) had little impact on ripple attenuation or F.B. GBW. The impact on surge voltage has not been investigated, but I would expect there to be an advantage in having more than half of the capacity on the filter output. On the other hand, ripple currents are higher on the first stage capacitors, and arguments could be made for shifting the balance that way.

**Current Feedforward**

The design of current feedforward circuits is also beyond the scope of this paper. Suffice it to say that output current sensing should probably be done with an air-cored "Rogowski" inductive pickup coil for minimum cost and insertion impedance. The coil would be physically located between or around the converter output conductors, just past the final output capacitor. The output of the pickup coil would be integrated (perhaps with a simple R-C integrator) and added to the current reference in the current mode control circuit, such that the current command is automatically adjusted for output load current changes.

**Feedback Design Considerations**

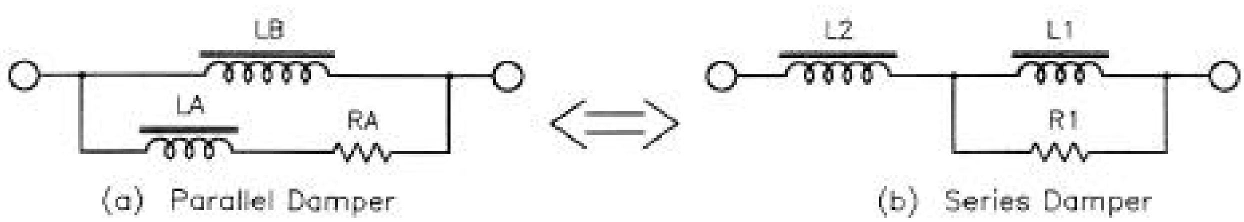
The feedback GBWs required for the filters discussed are much higher than normally used. The switching cell transit delay is largely unavoidable (and was taken into account in the modeling), but all other parasitic delays must be minimized, or at least held to a small fraction of the switching period.

If op-amps are used in the feedback loop, remember that delay increases with closed loop gain. High GBW amplifiers operating with minimum gain (near the unity gain crossover frequency) are recommended. If high gain is necessary, it is beneficial to cascade two or more amplifiers at lower gain [4]. I have designed my own amplifiers from discretes to achieve very high bandwidths (although moderate gain), with low power consumption and operation from output voltages as low as 2V [3].

Opto-isolators are generally much too slow for feed back across an isolation boundary, even using cascade followers for minimization of Miller capacity effect. Pulse transformers are a recommended approach.

**Converter Topology Options**

For reasons of simplicity and familiarity, the single ended buck converter with a transformer active reset was chosen for the prototype, but any other current mode controlled buck converter should work as well. Boost and buck-boost convert-




$$LB = L1 + L2 \qquad LA = \frac{L2\,(L1 + L2)}{L1} \qquad L1 = \frac{(LB)^2}{LA + LB} \qquad L2 = \frac{(LA)(LB)}{LA + LB}$$

$$RA = R1\left(\frac{L1 + L2}{L1}\right)^2 \qquad R1 = \left(\frac{LB}{LA}\right)^2$$

**Fig. 18 Parallel to Series Inductive Leg Damper Transformation**

ers have two drawbacks: the response time delay of a right hand plane zero, (although this effect is minimal when operating below critical current), and their current mode control response is not linear.

There are also a few design alternatives which reduce the ripple requirements, and are worth considering:

1) Two stage preregulator-converters , where converter output is nearly pure DC to start with, having only a commutation drop to filter [3].

2) Multi-phase converters, which provide a first order cancellation of inductor ripple currents, particularly at the fundamentalfrequency.

## CONCLUSIONS

It has been found possible to design high performance filters using MLP capacitors at low voltages and high conversion frequencies. Physical design and layout will be quite demanding, but then this is true of all aspects of high frequency converter design, and not peculiar to filters using MLP capacitors.

Some of the filter configurations and techniques developed may be useful in other applications where high ripple rejection is required while filter capacity needs to be minimized. Common and normal mode EMI filters for the AC input of switchmode supplies is one area, particularly when switching Unity Power Factor “front ends” are used. Another area of application might be for DC-AC inverter or switchmode audio amplifier output filters, where large filter capacities cannot be tolerated.

## REFERENCES

Other relevant papers by the author are:


[1] B. Carsten; “Slow Wave” Transmission Line Resonance Phenomena In Monolithic Ceramic and Stacked Foil Plastic Capacitors; Proc. of HFPC ‘91, Toronto, Ontario, Canada

[2] B. Carsten; Design of Multi-stage Filters For Use Within Wideband Control Loops; Proc. Of HFPC ‘92, San Diego, CA

[3] B. Carsten; A Low Voltage Schottky for High Efficiency VLSI Power Supplies; Proc. Of PCTM ‘84, Paris, France

[4] B. Carsten; Design Tricks, Techniques, and Tribulations at High Conversion Frequencies (tutorial); Proc. Of HFPC ‘87, Washington, DC; Proc. of PCI ‘87, Munich, Germany

# Follow Up Report and Initial Test Results for the 48V to 3.3V, 60A, 1MHz Prototype DC to DC Converter

Developed by:
Bruce Carsten

For:
ITW PAKTRON

Research to develop suitable filter designs using Multi-Layer Polymer (MLP) capacitors in low voltage, high frequency converters was funded by ITW Paktron, with the results of theoretical investigations and computer modeling reported in a paper given at the HFPC '95 conference[1]. A "proof-of-concept" 1MHz, 48 VDC to 3.3 VDC, 60A converter prototype was designed and built using one of the more promising filters (Figure 8 in[1]). The prototype was not ready for testing by the conference date, but has now been sufficiently completed to allow preliminary testing.

## PROTOTYPE CONVERTER CIRCUIT

A block diagram/simplified schematic of the prototype converter is shown in Figure 1. The power and drive circuit is complete as shown, but the control logic is still under development and somewhat proprietary, and is shown in a simplified form. The power circuit topology is that of a single ended forward converter using active transformer reset[2], with a three stage L-C output filter with inductive leg damping.

Dynamic current mode control is achieved on the secondary side by integrating the main output inductor voltage, with peak current limiting performed on the primary side. A primary side clock is used to turn on the main switches, with operation at a maximum 80% duty cycle. The duty cycle is shortened by turn off commands from the peak current sensor, or transformer coupled pulses from the output voltage regulator on the secondary side.

The voltage regulator has about a 100 KHz gain BW, with "output current feed-forward" to improve response to transient loads. The control and feedback circuit appears generally stable, but is unstable under some line and load conditions, particularly below critical current in the main inductor (<30 A out).

This partial stability with the high gain-BW feedback loop around a three stage filter indicates that the approach is basically valid, but there were insufficient resources left to track down the problems at this time. Efficiency and ripple data were taken by operating the DC-DC converter "open-loop".

## OUTPUT AND INPUT RIPPLE

Contrary to the computer model, the output ripple was found to be at a minimum when the resonant notching inductor (L4 in the schematic) was not used. Without L4 the output RMS ripple was about 7 or 8 times lower than predicted by the computer model, but with a much stronger relative harmonic content, as shown in the scope printout in Figure 2. The ripple waveform is closer to what would be expected with L4, granting the ideal capacitors (with ESR only) assumed in the model.

The MLP capacitors are not ideal in the 1-20 MHz frequency range, however, as can be seen from Figure 1 in the paper[1]. Transmission line resonances (and perhaps some ESL) cause various impedance peaks and dips in this frequency range, which can largely explain the difference between modeled and actual ripple.

The first capacitor impedance notch at about 1.4MHz causes the impedance at 1MHz to be about 1/2 that of the ideal capacitor. With three stages, this does much to explain the 8:1 lower RMS ripple than modeled. On the other hand, the much higher than ideal impedance at the harmonics of the conversion frequency (eg. 3 MHz, 5 MHz, etc.) also explains the high harmonic content of the output ripple when a nearly sinusoidal ripple would be theoretically expected.

Changing the orientation of the capacitor plates to normal to the mounting surface should eliminate the transmission line resonances, but would not affect the residual ESL of about 460pH. This should improve the harmonic rejection considerably, although the "resonant notch" inductor L4 would now be required to reduce the fundamental amplitude. This may be explored in the future.

The output RMS and P-P ripple measured with the Tek 11402 digital scope under various line and load conditions is shown in Table 1 below. The output ripple waveform under all of the tabulated conditions was very similar to that in Figure 2.

The modeled input ripple at 48 V and full load (approximately worst case conditions) was 8mV RMS and 22mV P-P,

| | | Table 1 | |
|---|---|---|---|
| V in | Iout | Output Ripple | |
| VDC | ADC | mV RMS | mV P-P |
| 36V | 40A | 1.43 mV | 9.94 mV |
| 48V | 40A | 1.92 mV | 9.86 mV |
| 48V | 50A | 1.94 mV | 10.5 mV |
| 48V | 60A | 1.98 mV | 11.3 mV |
| 60V | 40A | 2.07 mV | 9.63 mV |
| 60V | 50A | 2.13 mV | 10.7 mV |
| 60V | 60A | 2.13 mV | 10.8 mV |
| 72V | 40A | 2.31 mV | 9.95 mV |
| 72V | 50A | 2.42 mV | 11.1 mV |
| 72V | 60A | 2.43 mV | 11.0 mV |

while the measured ripple (see scope printout Figure 3) was 7.3mV RMS and 45mV P-P. Although the modeled and measured ripple voltages are in reasonable agreement, the ripple waveform is not. Capacitor resonances similar to those in the output filter again appear to be at work, as shown by the low fundamental and high relative harmonic content evident in the input ripple waveform.

## CONVERSION EFFICIENCY

A reasonable effort was made to keep the conversion efficiency high, although this was not the principle goal of the prototype. The measured efficiencies (including 1.33W logic and FET drive losses) are plotted in Figure 4, where they are seen to be in the 80-85% range over most operating line and load conditions, which is respectable for a 3.3V output, 1MHz converter.

## PROTOTYPE CONSTRUCTION

Some of the details of the output circuit construction may be gleaned from the working sketches shown in Figure 5, and are noted in the list below. Engineers desiring more information on MLP capacitors should contact ITW Paktron, while those interested in further converter design and construction details should contact Bruce Carsten.

**Features and Advantages of the Prototype**

1) All capacitors above 100nF are Multi Layer Polymer (MLP) type; no electrolytic capacitors are used in the converter.
2) The basic converter topology is a half wave forward converter with a transformer active reset circuit[2].
3) The 3.3V, 60A output has only 60uF of total capacity in a three stage L-C filter with inductive leg L-R damping. The worst case measured output ripple (at high line and full load) with a 20MHz BW was 2.5mV RMS (12mV P-P).
4) The input "Pi" filter has 11uF of total capacity, including the R-C capacitive leg dampers. The measured worst case input ripple (48 VDC in, full load) with a 20MHz BW is 7.3mV RMS (45mV P-P).
5) T1 and L1 share a common flux return path where the AC fluxes partially cancel, reducing core mass and losses.
6) Essentially two copper strap "bus bars" form all output side high current conductors, including inductor and transformer secondary windings and all interconnections.
7) The output rectifiers form two opposite sides of the T1 rectangular secondary winding, with the output + and - bus bars forming the other two sides.
8) L1 has a distributed gap "powdered iron" core (Micrometals "6" material) under the winding, with a high permeability ferrite return path. This eliminates external magnetic "fringe" fields, and forces AC currents to be uniform across the copper strap winding to reduce eddy current losses. Core losses are not excessive even at high line (about 1W), despite the 90 A P-P ripple current at 1MHz.
9) L2 and L3 (also with powdered iron cores under the winding with ferrite return paths) have tightly coupled "secondary" winding, allowing damper resistances to be in the "ohm" range instead of tens of miliohms.
10) Output capacitors C1-C6 are soldered to low thermal mass copper foils, which nevertheless form sub-nanohenry connections to the output copper bus bars.
11) Current in the T1 secondary and paralleled output rectifiers is "forced" to be uniform by the 6 turn copper strap solenoidal primary winding.
12) Main transformer T1 is gapped to lower magnetizing inductance, avoiding "reverse core saturation" with a relatively high leakage inductance and the use of the transformer active reset circuit[2].
13) High speed "AC" series CMOS is used for maximum logic speed, with discrete level shifting gate drive[3] to interface between the 6V logic and the 10V power FET gate drives.
14) A low insertion impedance (<2 nH) "Rogowski" coil provides output dynamic current sensing, allowing current feedforward to be used to improve transient load response significantly.

## REFERENCES


1 B. Carsten; "Optimizing Output Filters Using Multilayer Polymer Capacitors in High Power Density Low Voltage Converters"; Proc. of the IOth HFPC Conference, 1995, San Jose, CA.

2 B. Carsten; "Design Techniques for Transformer Active Reset Circuits at High Frequencies and Power Levels"; Proc. of the 6th HFPC Conference, 1990, Santa Clara, CA.

3 B. Carsten; "Design Tricks, Techniques and Tribulation at High Conversion Frequencies" (Tutorial); Proc. of the 2nd HFPC Conference, 1987, Washington DC.; Proc. of PCI '87 Conference, Munich, West Germany.

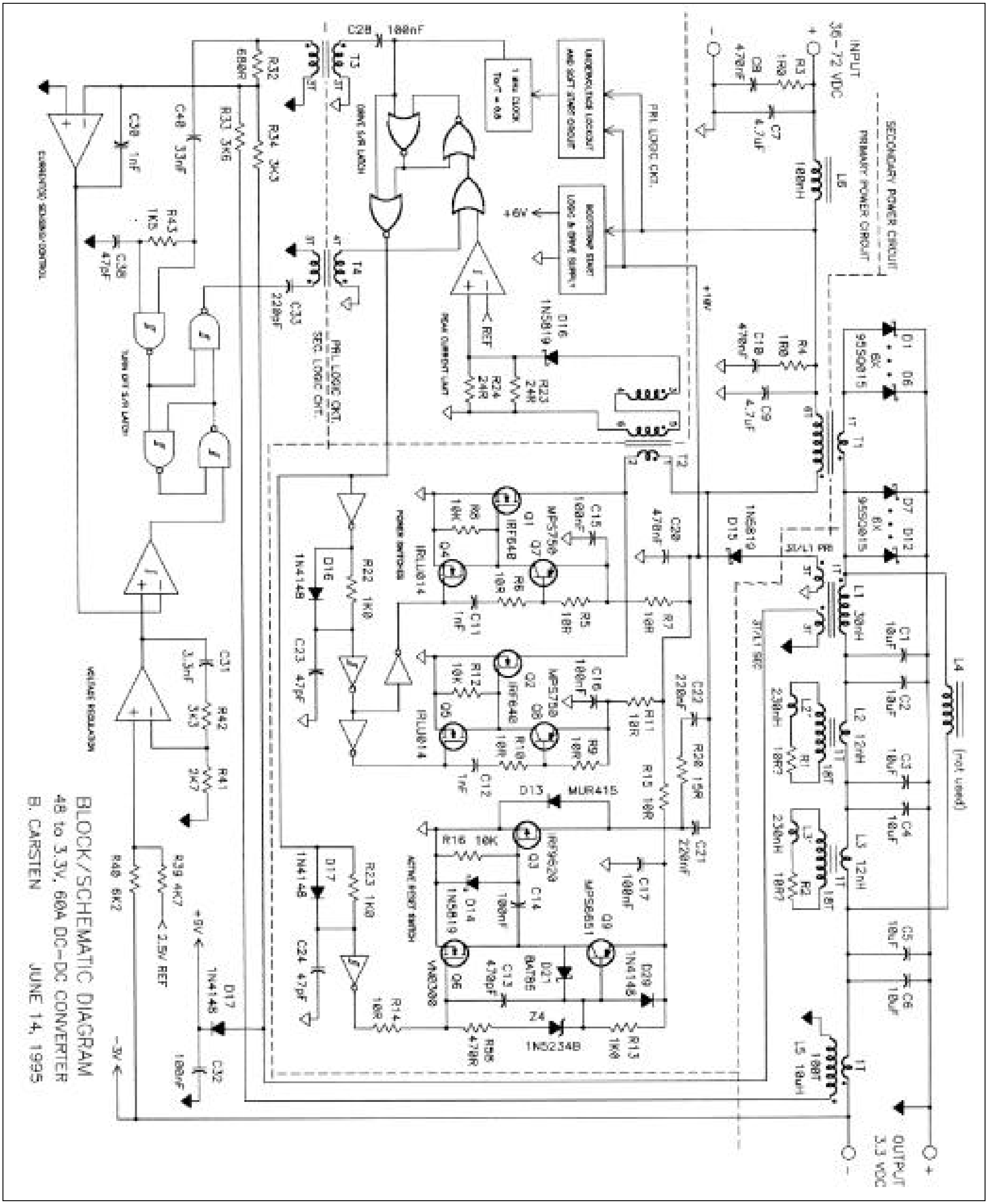



**Fig. 1 Block/Schematic Diagram. 48 to 3.3V, 60A DC-DC Converter**

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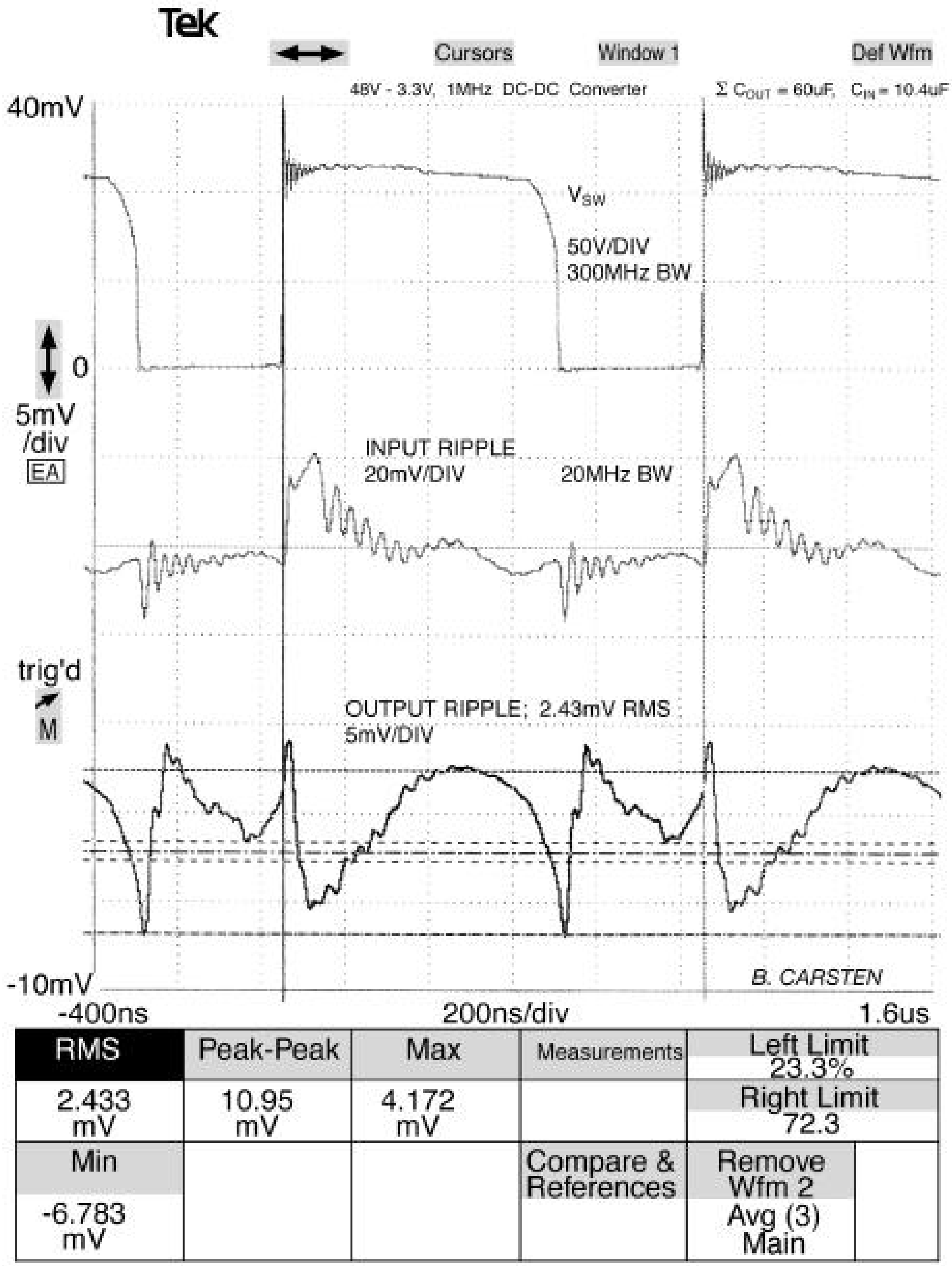


**Fig. 2 Worst Case Output Ripple**

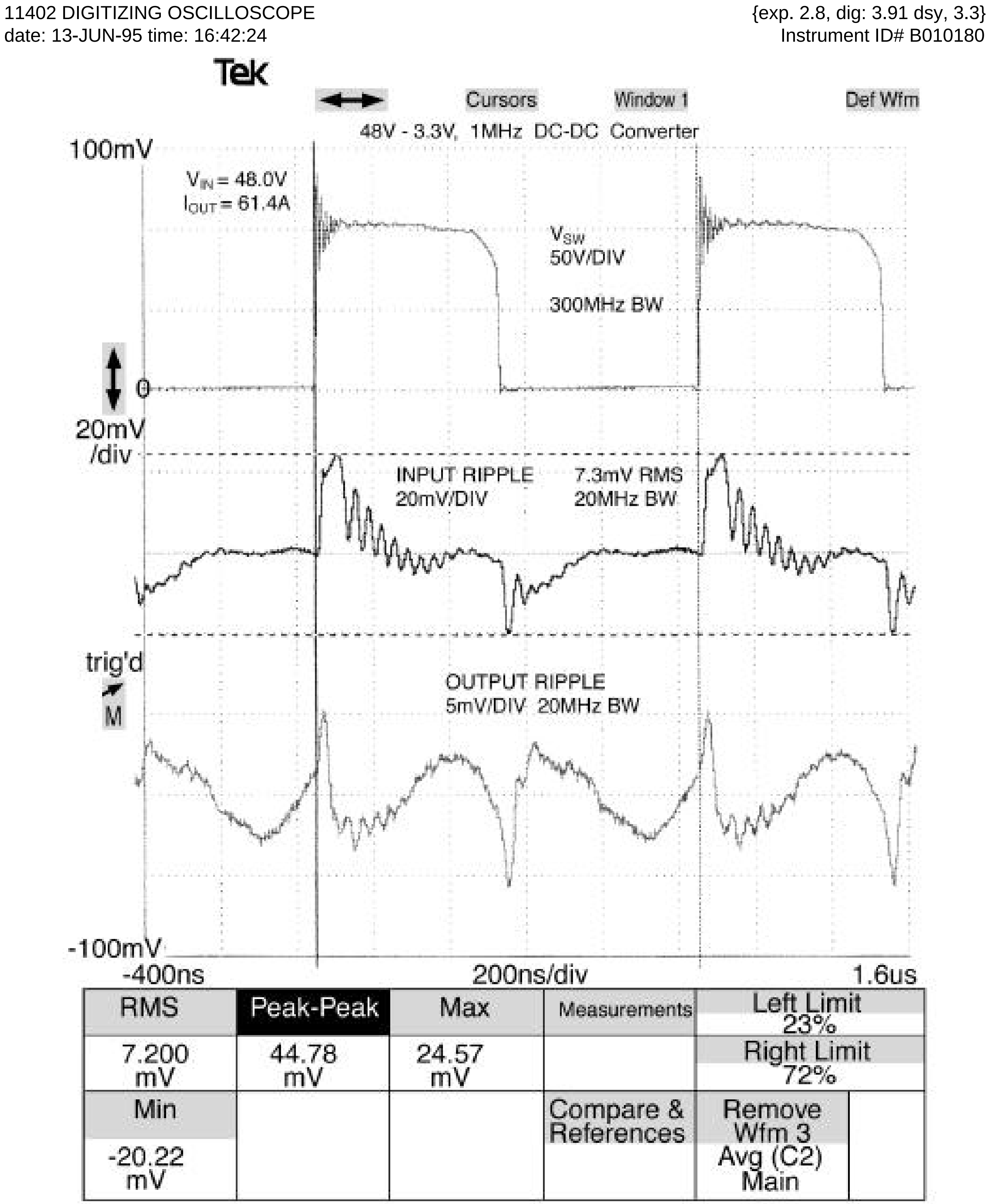


**Fig. 3 Worst Case Intput Ripple**

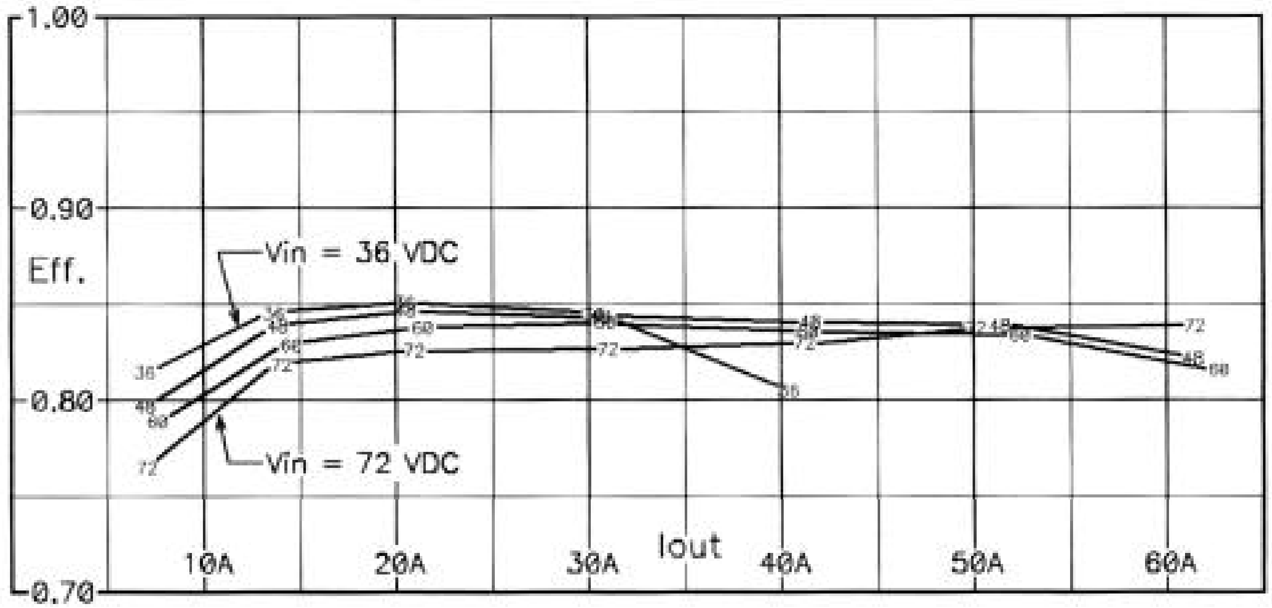


**Fig. 4 Effieciency vs Output Current & Input Voltage**
**48VDC to 3.3VDC, 60A, 1MHz DC-DC Converter**

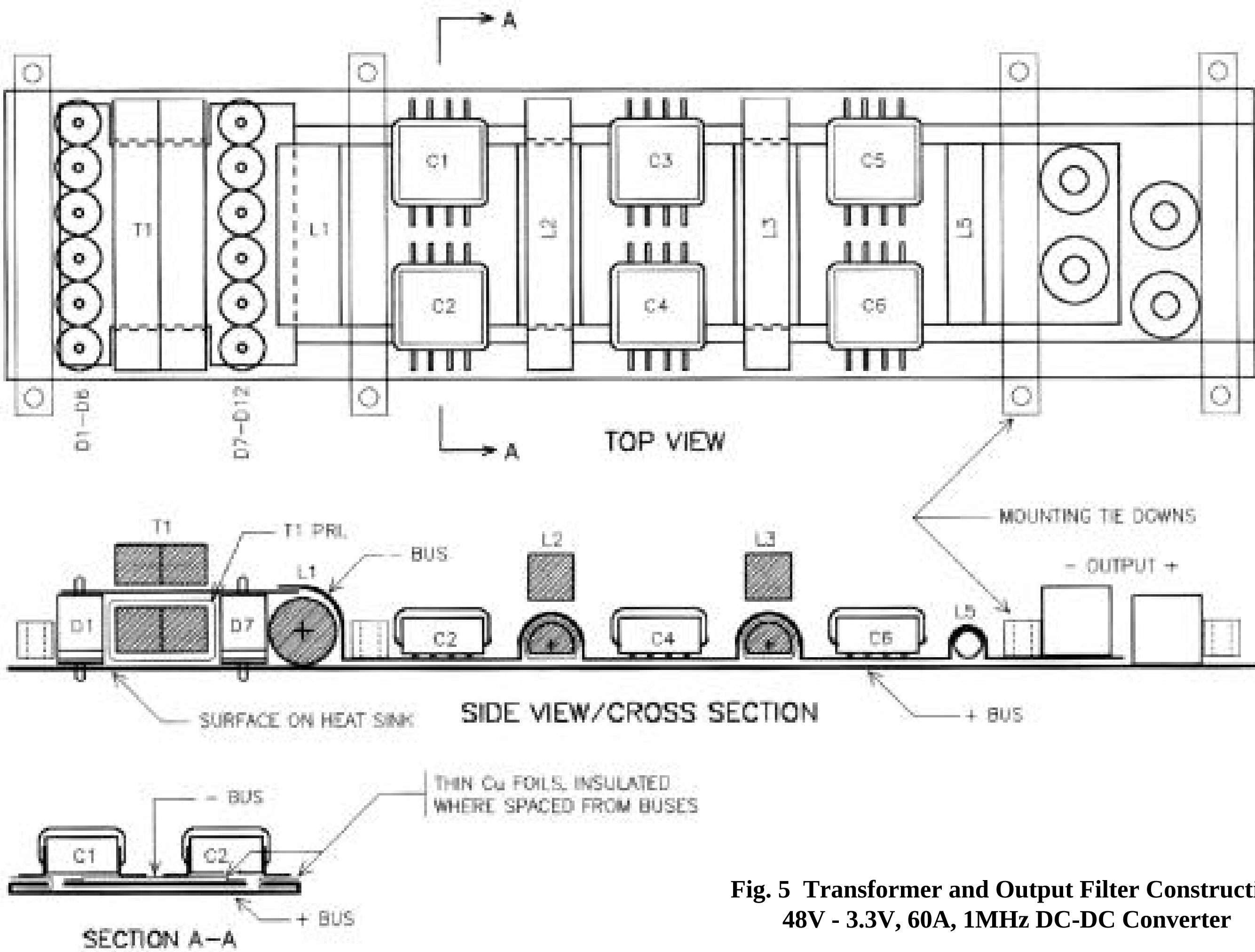


**Fig. 5 Transformer and Output Filter Construction**
**48V - 3.3V, 60A, 1MHz DC-DC Converter**