Putting Profit Into Power Electronic Products with Digital Control

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Abstract—

This paper discusses challenges and opportunities in making commercially successful power electronic products that incorporate digital control.

I. Background

At the risk of making a trite observation, the last two decades of advances in microcontrollers, processors, and programmable logic have opened up tremendously exciting possibilities for enhancing the performance, applicability, and economy of power electronic supplies and drives. Adaptive controllers, parameter estimation, and sophisticated control algorithms, while not new concepts, have become much more economically reasonable to implement. Power supplies and drives do not have to serve simply as power transformers. A power supply controller can beneficially affect the overall performance and dynamics of a complete system, including servomechanical loads, discharge illumination, high-end central processing units, and other loads with significant commercial and industrial relevance.

The often recited benefits of programmable digital logic include: flexible reprogramming during design and in the field; the possibility to incorporate value-added functions such as sophisticated user interface; ability to implement adaptive, multi-input/multi-output, and nonlinear control strategies; easy compatibility with optical isolation; and the ability to execute diagnostic fault detection and reconfiguration codes as background processes. Do these benefits offer real commercial value to practical power electronic products, or are they academic curiosities? The "cons" of digital control can be significant burdens to the designer and manufacturer. In comparison to the bread-and-butter analog control techniques and components that serve much of the power electronics community, digital control introduces a host of new issues: the effects of quantization and approximation in arithmetic; the challenge and expense of including code development, revision, and correction as additional design steps; the hardware expense or perceived expense of "going digital" in comparison to using similar bandwidth analog components; and the challenge of developing and using valid circuit models and control schemes for a digital approach when our community has worked so hard and so successfully to build continuous-time (CT) models.

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The open marketplace will continue, no doubt, to provide Darwinian answers to the questions of when and in what applications digital control will make practical sense for power electronics. The prescient (or the clever or lucky) who are able to sense the right time and the right application for any technology are more likely to develop products that dominate future markets. As a technology, digital control of power electronics is no exception.

The purpose of this paper is to provide a speculative glimpse, based on historical and current trends, of commercially valuable applications of digital control in power electronics. The next section reviews a few basic issues of control and digital control specifically. This review is intended as a short reminder of some of the differences between continuous and discrete time systems, and how these differences might influence a designer. Following this review, the remaining sections examine specific applications, opportunities, and speculations regarding digital control in power electronic systems. The recent Special Issue on Digital Control of the IEEE Transactions on Power Electronics serves as a foundation for this discussion [1].

II. Review of Discrete-Time Systems

Let's start with a "basics" review and comparison of continuous and discrete time systems and controls. For any specific plant, the choice of continuous or discrete time control implies a choice in hardware and modeling techniques. An exhaustive tutorial could

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occupy one or more full papers. This section highlights some of the "new" technical issues that arise when contemplating digital control starting from a good familiarity with classical analog control. The nature of the plant and the customer specifications ultimately determine which design choices will prove economically and technically feasible. This review sets the stage for the remainder of the paper – speculating where digital control may prove economically beneficial in future power electronic systems.

There are at least two reasons why a designer might be confronted with a digital control design problem. The first is because the plant is innately best described in discrete time (DT). An everyday example of this situation is a mortgage with regular monthly payments [2]. In this example, it is natural to define a discrete time variable, $P[n]$, the unpaid principal owed to the lender after the n^{th} monthly payment of p dollars has been made. Given a monthly interest rate r equal to the annual percentage rate divide by 12, a DT difference equation can be developed that describes the dynamics of the principal:

$$P[n+1] = (1+r)P[n] - p \qquad (n \geq 0) \tag{1}$$

This equation is the start point for a number of different dynamic analyses. For a fixed term mortgage, e.g., a 30 year (360 month) mortgage, the requirement $P[360] = 0$ can be used as a constraint to study the effect of different plant time constants (determined by the interest rate, r) on the drive p necessary to meet the final value in the required time. A control system arises when we servo the monthly payment p to retire the mortgage early. An error or difference between the actual remaining principal and the desired remainder might cause us to increase p to drive the system to $P = 0$ more quickly (improved transient response). This familiar request for improved response time comes with an equally familiar requirement for more "drive" capability – in this case, the ability to make a larger monthly payment than the minimum required. Instability can arise when the monthly payment p is insufficient in the face of the interest rate r to prevent unbounded growth in the owed principal P.

There are cases where power converters can be most conveniently modeled directly in discrete time. A power-factor correcting utility interface, for example, will typically take voltage-loop control actions at most twice per utility line cycle (and generally less often than this [3]) to avoid disturbing the input current waveshape. These converters are well-described by DT difference equations, and digital control is an obvious choice [4], [5].

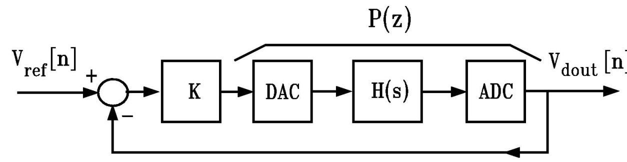

Fig. 1. Discrete-time feedback loop.

The second reason why a designer might face a DT control design problem arises when some aspect of a digital controller, e.g., adaptability, is necessary to meet customer performance demands. In this case, the control hardware imposes the need to deal with DT modeling. This situation is illustrated in Fig. 1, which shows a CT plant with transfer function $H(s)$ embedded within a single-input, single-output (SISO) feedback loop. A zero-order hold or digital-to-analog converter (DAC) and a sampler or analog-to-digital converter (ADC) are used in this example. The DAC converts DT control commands to CT inputs to the plant. The ADC samples the CT output of the plant to create DT measurements for the controller. A series compensator is included in the control loop, in this case, a proportional gain K. The unity feedback configuration suggests a regulator, typical for many power supply design problems. Taken as a group, the DAC, plant $H(s)$, and the ADC form a single DT block with a control input and a sampled output. The dynamics of this "macro-block" can be modeled by a z-transform, $P(z)$. Just as the Laplace transform can be thought of as a frequency-domain transform or operator calculus where the frequency variable s serves as a place-holder for a time derivative, the z-transform is a frequency-domain transform where the frequency variable z denotes a time difference.

The system in Fig. 1 can be used to understand some of the twists that arise when contemplating a DT feedback loop given a CT design background. To make the example concrete, pick a simple system for

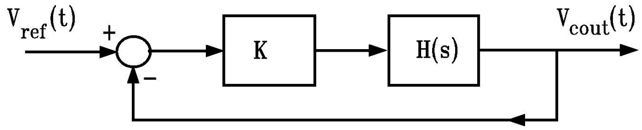

Fig. 2. Continuous-time feedback loop.

the plant, e.g., an RC divider circuit with

$$H(s) = \frac{1}{RCs+1} = \frac{1}{\tau s+1} \tag{2}$$

where τ is the RC time constant. In a power electronics application, a linear, time-invariant (LTI) plant model $H(s)$ might only be developed after more modeling effort, e.g., averaging and small-signal linearization. The DT transfer function $P(z)$ of the macro-block is related to $H(s)$ by a step-invariant transformation [2]. A DT unit-step input to the DAC results in a CT unit-step input applied to the plant, $H(s)$. For a unit step into the DAC, the output of the ADC must be samples of the CT unit-step response of $H(s)$. This required relationship, that the DT step-response of $P(z)$ must be samples of the CT step response of $H(s)$, is imposed by the DAC and ADC hardware used to construct the feedback loop. Given $H(s)$, $P(z)$ can be determined by exploiting this mathematical relationship. Digital control textbooks typically provide tables that relate common $H(s)$ plant models to their step-invariant transform models, $P(z)$ [6]. In this case, for the RC plant,

$$P(z) = \frac{1-\lambda}{z-\lambda} \tag{3}$$

where

$$\lambda = e^{\frac{-T}{\tau}} \tag{4}$$

and T is the sample period. Let's compare the behavior of this DT feedback loop to a CT feedback loop, as shown in Fig. 2.

Both Figs. 1 and 2 employ a subtractor and gain block K to implement the feedback loops. In the digital controller, these would be implemented as either dedicated digital logic (FPGA, discrete logic) or as lines of code on a microprocessor – both relatively expensive choices compared to the op-amp(s) that might be used to create Fig. 2.

How might a designer analyze the performance of these two systems to evaluate the "three pillars of customer satisfaction": transient response, steady-state tracking, and stability? System performance is typically evaluated using one or more of four methods for LTI SISO feedback loops (DT or CT): pole location methods like direct solution or root locus; Nyquist plots; Nichols analysis; or Bode plots. All four of these methods examine the loop gain to determine closed-loop performance.

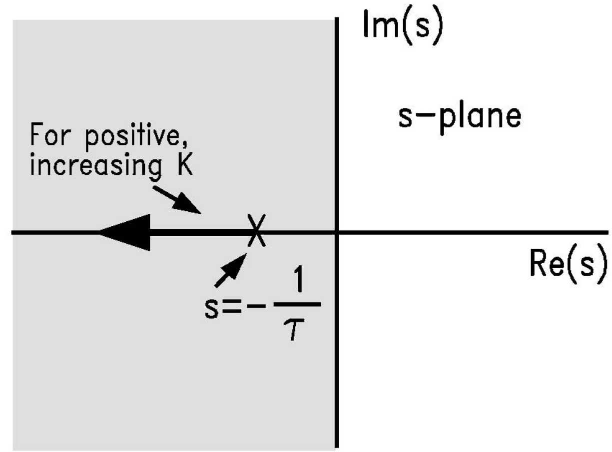

Fig. 3. S-Plane root locus.

Figure 3 shows a root-locus plot for the CT feedback loop in Fig. 2. This plot indicates the closed-loop pole location as K varies from zero to huge positive values. Ignoring real world issues like saturation and unmodeled dynamics, the root-locus plot supports the usual sophomore observation that "gain is good." As the gain magnitude increases, the closed-loop pole location moves deeper into the left-half plane (shaded region indicating stability). Transient response improves, the system is always stable, and larger gain K increases the loop gain at all frequencies, improving steady-state tracking.

Any method available for generating a CT root-locus plot carries over for DT root-locus analysis without change. All the familiar methods, e.g., application of the root-locus rules, or direct solution in Matlab or by other computational means, will produce a proper root-locus plot for Fig. 1 given the DT loop gain

$$L_d(z) = \frac{K(1-\lambda)}{z-\lambda} \tag{5}$$

for this example. It requires the same effort to root a polynomial in s or in z. The loop gain is first order and the root locus, shown in Fig. 4, is qualitatively

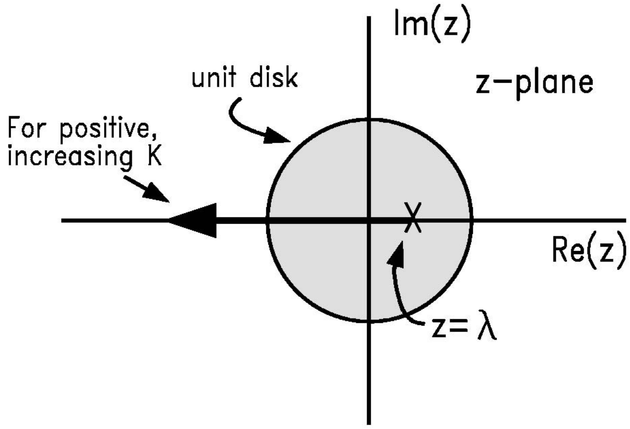

Fig. 4. Z-Plane root locus.

similar to Fig. 3.

Both root-locus plots, for low values of K, begin near the open-loop plant pole location: $s = -\frac{1}{\tau}$ for the CT system and $z = \lambda$ for the DT case. The time constant τ depends strictly on physical parameters and will always be positive for real resistors and capacitors. The z pole location λ will also be positive. It depends not only on the component values but also the sample period T. Stability in the z-plane means that the closed-loop pole must remain inside the unit circle, shaded in Fig. 4. Pole locations on the right-half real-axis inside the unit disk correspond to monotonically decaying transient responses, analagous to the closed loop behavior of the CT control loop. Locations on the left-half real axis inside the unit circle correspond to time domain transients that decay, but also oscillate! Pole locations outside the unit disk lead to transients with unbounded growth, that is, instability.

So, for our first order DT control system, the root-locus plot reveals interesting behavior as K increases: an initially stable first-order system that, for larger K, eventually oscillates and finally becomes unstable. This behavior is foreign to the CT loop and our intuition from first-order CT control loops. The DT control loop is "ballistic" in between sample instants. It exerts a control action at most every T seconds. As K grows while T remains constant, the drive to the plant increases in response to a given error. Larger and larger values of K lead to larger drives, allowing the system to "get away from us" at some point – the system response between sample instants becomes huge, with no correction until the next sample instant, leading to an oscillating instability as the controller "panics," driving the system between positive and negative extremes at each sample point. Note also that the designer now has two "handles" that control the system performance: the gain K and the sample period T. A larger sample period T is more economical in the sense that it lowers the controller's computation burden and permits the use of potentially cheaper DAC and ADC hardware. However, increasing T decreases λ, causing the destabilizing effects of excessive gain K to appear for lower values of K.

In a power electronics application, additional limitations on the sample rate might have to be considered. The sample rate must be selected to avoid aliasing ripple and high frequency disturbances. It must also be selected to avoid control actions that occur at a rate faster than reasonable for the modeling assumptions made during averaging and linearization [7]. These limitations can create both upper and lower bounds on the desirable sample period.

A little more thought is required to extend frequency domain methods like Nyquist, Nichols, and Bode plots to DT systems. The essential basis for all classical frequency domain performance analysis techniques is the Nyquist method. For a CT system, the Nyquist method invites the designer to plot the loop gain $L(s)$ in a real-imaginary plane as s varies around a contour. Although not a requirement, s is often chosen to vary on a "D"-shaped contour that includes the $j\omega$ axis and extends out over the entire right-half s-plane [8]. For this choice, the number of closed-loop system poles Z in the right-half s-plane can be determined as the sum $N + P$, where N is the number of times the loop gain plot encircles the -1 point, and P is the number of open-loop poles of the loop gain inside the D-contour. Generally, the design goal would be to have $Z = 0$ to yield a stable closed loop system. For our CT control system, the RC plant is open-loop stable, and $P = 0$. To create a stable closed-loop system, a designer would generally attempt to create a loop gain $L(s)$ such that N was also zero, i.e., no encirclements of the -1 point, leading to $Z = N + P = 0$ and stability.

Nichols and Bode plots are derivative of the Nyquist method, again providing stability and performance information by examining the frequency response of the loop gain. Bode plots are the most familiar and quickest graph for an engineer to generate. They are by far the most commonly used stability and performance analysis tool for CT control design. The common metrics of phase and gain margin are determined from Bode plots to evaluate stability and peformance. For minimum-phase, open-loop stable systems with simple loop gains, e.g., monotonically decreasing in magnitude and phase, the notions of gain and phase margin relate directly back to the Nyquist stability criterion. For these simple systems, gain and phase margin correspond to indications of how much the Nyquist loop gain plot can be expanded (gain margin) or rotated (phase margin) before causing a destabilizing encirclement of the -1 point. Even for CT systems, the notions of gain and phase margin become much trickier or even impossible to apply for very wiggly loop gains or in the presence of right-half plane singularities (despite the occasionally observed tendency to almost blindly rely on margins as indicators of stability and performance).

All of these frequency domain methods can be extended, with varying degrees of practical utility, to DT control analysis problems. The Nyquist method applies directly. In the case of DT systems, the D-contour would typically be replaced with an "O"-contour, a circle on the unit disk. For this choice, an open-loop stable DT loop gain would yield a positive P. To achieve a stable closed-loop, a designer would now need to generate a loop gain that produced a Nyquist locus with no encirclements of the -1 point, to ensure that $Z = N + P = P$. That is, we still seek no encirclements for our open-loop stable system, this time in order to ensure that the open loop stable poles remain in the "O"-contour in the closed-loop implementation.

Just as for CT systems, concepts of gain and phase margin could be developed that indicate the degree to which encirclements have been avoided. However, Bode plots are generally less easy to sketch quickly and intuitively for DT systems. The plots are periodic, and evaluated with an exponentiated frequency $z = e^{j\omega}$, as opposed to the more familiar $s = j\omega$. The effect of "standard" compensators may also be unintuitive on the DT Bode plot.

A classically trained engineer familiar with the powerful CT frequency analysis tools may therefore be uncomfortable working with DT systems. Our tendency to want to cling to these successful and familiar CT methods has led to the development of all sorts of *ad hoc* schemes that permit compensator design in CT, followed by a "final step" of mapping a CT design into DT with the prayer that the DT performance will be good enough. If it is not, the design process becomes an interative game of bouncing back and forth between CT designs and DT mappings. This game rarely results in a controller that produces the best system performance with the lowest sampling rate and computation burden.

"Modern" approaches to DT control employ state-space methods and eigenanalysis, which essentially boil down in many cases to direct pole placement and pole location for the closed-loop design. These methods can be computationally more intensive than the intutive design techniques built for CT systems. Of course, the computation technology that makes digital control a practical option also facilitates DT control system design. Nevertheless, for an engineer most familiar with CT design, there is naturally some resistance to adopting new design workflows to implement DT controls. Given the easy familiarity of classical CT controllers and the relative inexpense of CT control building blocks like op-amps, compelling economic reasons must be present to justify taking on a DT control problem in the power electronics environment.

III. Commercial Opportunity

Historically, programmable logic seems to make its way into practical power electronic and electromechanical systems when at least one of two conditions warrant the inclusion. First, some electrical and electromechanical components require complex control schemes to function at all in certain applications. These situations essentially demand the flexibility and sophistication of a microcontroller. Programmable logic is simply the easiest way to implement certain systems. Second, many applications require a finite-state machine or microprocessor to provide a desired user-interface or to minimize complexity for an operator. Digital control in such cases is a natural outgrowth of having a programmable logic device with "spare" computation power in the system architecture. It seems very likely that the well-known advance of microprocessor performance and the retreat of microprocessor cost are likely to continue in the near future. Many conceivable applications of digital control in power electronics are therefore already possible or soon will be. Whether or not these applications become profitable business opportunities depends on the presence of one or both of these stated conditions, and also on customer demand sufficient to make the incremental cost and design overhead of microcontrol an economically reasonable proposition. Two historical examples come to mind.

The first exemplifies a function-driven application of microprocessors. Despite the inconvenience of commutator brush maintenance, brushed DC motors dominated many position and velocity servo applications for roughly 80 years, either directly or as part of combined drives like Ward-Leonard systems [9]. This dominance stemmed from the fact that brushed DC machines can operate from two direct currents and are relatively easy to control. Blaschke's introduction of the concept of field-oriented control demonstrated that the complex dynamics of an induction machine could be reduced to a relatively simple description [10]. The "direct" approach to field-oriented control made it possible to bring the reliability of the induction motor to servomechanical applications, but compromised this reliability with the need to add field sensors in the machine air gap. The "indirect" approach to field-orientation, pioneered in publications like [11], used microprocessors to estimate the essential machine state variables, eliminating the requirement for field sensors. As with most engineering trade-offs, there was no "free lunch." The reliability of an overall drive system is generally enhanced by the use of a mechanically pristine induction motor without air-gap field sensors. On the other hand, the overall reliability of the indirect approach depends on precise knowledge of the induction machine parameters (resistances, inductances, etc.). The microcontroller can continuously update parameter estimates while the drive is operating. Microcontrolled ac motor drives have replaced dc machines as the dominant drive of choice in a huge number of applications, including variable speed drive systems in HVAC and people-movers like elevators.

A second, user-interface driven example of commercially relevant digital control arises in the case of contemporary distributed power architectures for telecommunication and computer systems. As noted in [12], "we are in the middle of a strong migration to distributed architectures for a variety of applications." In the 1980's, power electronics designers began a transition from more centralized power distribution in computers to systems with "board-mounted" power or arrays of point-of-load converters. It became economically feasible and desirable to include a microcontroller in many computer workstation power distribution systems to handle "sequencing." That is, a microcontroller might be included in a workstation just to switch on loads in a proper sequence. The microcontroller made it easier for a user to interact with a computer, both through automatic sequencing and also, for example, by updating a status and diagnostics display. Once the microprocessor made it way into these systems, it became natural to consider its use for additional purposes. Microcontrollers were pressed into service as voltage loop controllers, for example, in front-end ac/dc utility interfaces, providing higher bandwidth control and large-signal stability for unity-power factor (UPF) interfaces in distributed power architectures [4], [5]. In this case, higher bandwidth control and large-signal stability resulted from a reconsideration of the model of a boost-converter UPF in discrete time, an acknowledgement of the cyclical periodicity of the operation of the UPF interface. A similar observation can be made for many commercially available uninterruptible power supply systems.

There are at least eight application areas identified in the Special Issue [1] where digital control could have a direct impact on product value:

- Board-mounted power, dc-dc, point-of-load, and voltage regulation module (VRM) power supplies [13] (See also [12].).
- AC-DC converters for utility interfacing [14], [15].
- Flexible ac transmission for the electric utility [16].
- Uninterruptible power supplies [14].
- Illumination electronics [17].
- Electric drives for both rotary and linear machines in industrial, commercial and military applications [18].
- Power electronics for transportation systems and people movers [19] (See also [20]).
- High-fidelity audio applications [21].

How far will digital control spread into each of these areas? To what extent will digital control provide advantageous features that improve the profitability and customer satisfaction of power electronic products in these and other areas?

To find places where digital control will have technical and commercial relevance, look for applications where either discrete-time modeling and digital control essentially make the application possible, or where customers will demand or can be enticed to demand the features of a digital interface. If either of these two drivers exists in an application, then both might be able to be implemented economically now or in the near future. The next sections look at three of the eight application areas listed above for representative clues about where digital control will enhance future profits in power electronic products.

IV. Board-Mounted Power and VRMs

A flurry of activity in the 1980's and early 1990's demonstrated that arbitrary increases in switching frequency do not necessarily reduce overall dc-dc converter volume [12]. Nevertheless, unrelenting pressure to improve transient response is, in some cases, encouraging higher switching frequencies even in the absence of any appreciable increase in net power conversion density [22], [13]. Extremely high switching frequencies, high power densities, and snappy transient response may be possible at high efficiency from converters with different circuit topologies or different component technologies from the classical canonical-cell-derived and quasi-resonant and resonant converters that make up the core of point-of-load products in the marketplace today. Switch performance and magnetic core materials are currently key limiting factors in the efficient increase of switching frequency and power density.

Speculatively, we may find that at extremely high frequencies, micro-fabricated inductors or transmission lines with reasonable quality factors might be fabricated on-chip without the use of high-permeability magnetic materials [23], [24]. It has been suggested that micro-transmission lines can be arranged as delay lines to reflect or transmit pulses at high frequencies with timing and harmonic content suitable to replace some of the switches in a microfabricated power converter. Digital control may have a central role in such a converter. A circuit dependent on the tuning of passive transmission lines may need parameter estimation to adjust and tune switch frequencies for proper operation. It may also be convenient to integrate digital control along with the converter.

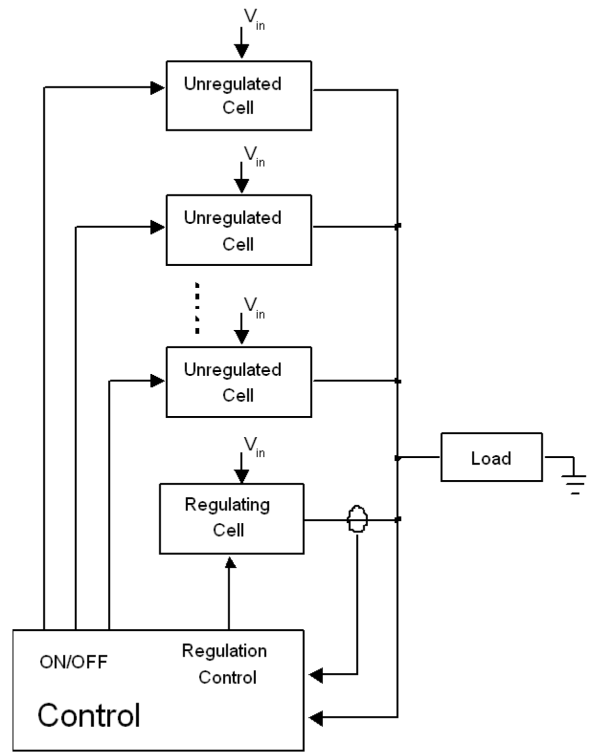

Fig. 5. High-frequency converter architecture (courtesy of Professor David J. Perreault).

Digital control may be essential in future, high-frequency miniaturized power supplies for at least three reasons. First, high frequency "on-chip" or "on-module" power supplies may employ power conversion architectures that are most naturally controlled based on a digital model. Second, as voltage rails decrease in many digital and mixed-mode loads, it may become extremely difficult to implement analog control with adequate noise rejection. A significant advantage could prove to be the noise margin or noise rejection capability of a digital controller. Third, control algorithms for very-high-frequency conversion architectures may require precise knowledge of component value or tuning. In this case, digital estimation may be a critical enabling technology for controlling these power supplies [25].

The converter architecture discussed in [26] and shown in Fig. 5 (courtesy of the authors of [26]) could conceivably be used to create compact, RF frequency converters. These converters could be suitable for integrated fabrication using batch methods, and could be integrated with digital or mixed-mode integrated circuit loads or as part of modules. In the scheme discussed in [26], the overall converter architecture consists of a bank of high-frequency, unregulated cells that can each process a quantum of power. These cells switch at a sufficiently high frequency to employ air-core, miniature magnetic energy storage components. The authors of [26] are, for example, exploring the implementation of each cell based on a self-oscillating Class-E converter. Each cell is designed with an output impedance or "droop" characteristic that permits current sharing. The control circuit sends a digital ON/OFF signal to each unregulated cell, engaging enough cells as needed to provide the bulk of the load power. A regulating cell serves as a "vernier" to complete output voltage or current regulation.

This circuit presents opportunities for digital control. The ON/OFF cell selection could certainly be handled as a digital task, especially given the need for hysteresis to avoid chatter as "cell boundaries" are crossed. The control of the regulating vernier cell could be either analog or digital. An analog controller with very low supply voltage rails could be developed using sub-threshold circuit design techniques. Making such circuits work is likely to be difficult in a switching noise environment. An analog controller could also conceivably be developed using conventional analog components and a higher voltage bus (if available) to power the control circuitry. This approach demands the availability of additional high voltage power supplies. Also, to bring measured signals into and out of the control circuitry, low-noise amplifier (LNA) front-ends and attenuating drivers might be required. The construction of these LNA components could prove economically unattractive. Increasingly, there appears to be a trend to employ digital control for low-voltage converters, particularly those in high-noise environments, e.g., driving digital or mixed-mode loads [25], [27], [28].

The architecture shown in Fig. 5 also offers the possibility of enhancing overall system performance by providing fault tolerance and dispersal of heat generation. Again, digital control may be essential to squeeze the most from these features. A digital controller performing parameter estimation could conduct sensorless tracking of thermal conditions. It could also keep track of module health and reconfigure around damaged cells to provide partial operating capability.

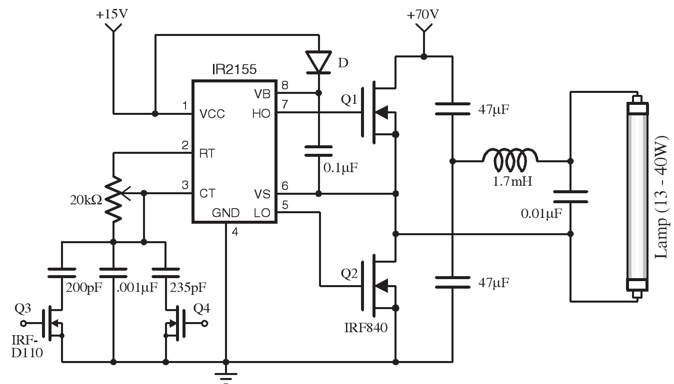

Fig. 6. Lamp ballast circuit.

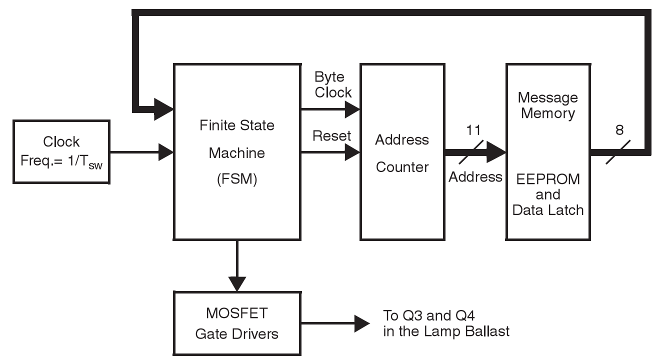

Fig. 7. Transmitter block diagram.

V. Discharge Illumination

A less speculative example may be found in the case of discharge illumination. A fluorescent lamp ballast is essentially an ac-ac converter that draws power from the utility and provides striking and operating ac waveforms to a lamp tube or tubes. There has been steady pressure over the last 30 years to increase the operating frequency of fluorescent lamps substantially over the 60 Hertz used in traditional iron-core ballasts. Fluorescent lamps operate more efficiently at moderately higher frequencies (in the tens of kiloHertz) because the faster drive prevents cooling of the lamp plasma and the need for cycle-to-cycle reignition of the lamp.

Solid-state converters in ballasts have become an important commodity item. They produce high frequency waveforms that improve lighting efficiency. They permit direct control of the lamp voltage and current waveforms, enabling dimming, crest-factor control (which enhances bulb life), and filament warming during starting. Microcontrol is a natural addition to a solid-state ballast, permitting luminaires to respond to control signals from room occupants or facilities managers, e.g., the digitally addressable lighting interface (DALI).

Once a ballast contains a solid-state drive and a microcontroller, a host of new product opportunities arise. In [29], a solid-state ballast is introduced that uses a microcontroller not only to control the lighting but also to modulate the current in the lamp to provide optical communication as shown in Figs. 6 and 7, from [29]. A "talking" ballast that can send optical information to a wearable receiver becomes an information node that can function like a "satellite" in an indoor GPS system. Occupants in a building can receive messages from the lights keyed to their specific location. For example, the blind can be guided around shopping venues by following signals from the lights [30]. In this application, digital control enhances the ballast, making it far more than a commodity item, bringing it into a building's information network.

In [17], the authors present another illumination example where digital control enables efficient operation of metal halide lamps. Inefficient low frequency operation of metal halide lamps is commonly practiced to avoid acoustic resonance. A digital controller is developed in [17] that employs a modulation pattern that thwarts acoustic resonance and permits higher frequency operation, in principle improving overall lamp efficiency.

VI. Electric Drives and Motion Control

Microprocessors and digital control made an early appearance in power electronic drives for motion control systems. The use of digital processing in this arena has grown unabated. This trend is guaranteed to continue, especially as increases in processing power and communication bandwidth continue while costs decrease.

Some hints about where motion control system designs may be headed can be found by looking at current market leaders. Adept Technology, Incorporated, for example, makes a line of flexible servo kits and multi-axis motion systems for industrial robotic manufacturing. Their product line skirts, and sometimes sets, the edge of applications of digital technology in motion control. Their SmartAmp system, for example, is fully digital directly to the gate drive level [31]. A DSP and FPGA combination is used in each SmartAmp module to control every aspect of converter operation. The DSP implements the motion control algorithms, and the FPGA handles direct PWM generation and unity-power-factor interfacing to the utility.

The fully digital approach gives the SmartAmp module several valuable advantages. Each module controls one axis of a multi-axis robot, which can be flexibly arranged according to customer requirements. Full digital operation means that it is easy to coordinate the activities of each module with a minimum of connecting wires. An IEEE 1394 bus connection can be made between each module, permitting high speed data exchange. A maximum of 13 conductors runs between each module, replacing 120 wires (for power, encoder information, status, etc.) in an earlier system. This reduction in wire bundling improves reliability in the overall robot and eases the critical problem of mechanical lead dressing. Because the modules are fully digital to the gate drive level and because they are interconnected by a high-speed data bus, power switching can be coordinated across modules. That is, a shared clock signal can be passed between modules and used to coordinate switching activities to minimize the generation of conducted EMI.

The SmartAmp module is probably close to the limit of affordable technology in its particular niche. Technology like that found in these modules is essential to meet the strictest harmonic standards and to absolutely minimize wiring and maximize kit flexibility. It is relatively expensive, and at the edge of achieving acceptable stator current bandwidth with an economically acceptable digital controller. Where regulatory requirements are lower, other, less expensive amplifier designs are available that offer different trade-offs. The "Amplifier-in-Base" (AIB) design found in the recent Cobra robot, for example, uses analog current loops at the innermost current generation stages for the motor stator. This potentially less expensive architecture still uses a DSP and an FPGA to generate reference signals for the analog subsystems.

It remains to be seen how quickly DSP and FPGA technology will advance in capability and decrease in expense to the point where all-digital designs consistently become the obvious choice for motion control. The stringency and extent of regulatory requirments on conducted EMI, for example, will effect this decision. Digital technologies for randomizing or spreading EMI may become very attractive in comparison to analog filtering in the face of very strict regulatory requirements. Significant cost reductions in large (in excess of 200,000 gates), high speed FPGAs wil also accelerate the penetration of digital control deeper into servo control systems.

VII. Discussion

The creativity of our colleagues exemplified in the recent Special Issue speaks convincingly to the notion that there is a growing role for profitable applications of digital control in power electronics. There are more hints of the importance and utility of digital control in the Special Issue than can be covered in this paper. Hopefully, the abbreviated review conducted here provides some provocative criteria for evaluating the likely success of potential applications of digital control. A careful review of current power electronic applications with an open mind about digital control may also suggest new research and development possibilities for digital control technologies. The expanding availability of higher capability digital tools makes this a very exciting time to be designing power electronics.

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