## CONTENTS

## 1. TERMINOLOGY

Reading a specification for an inductor is considered a simple task, however, there is often some confusion even over the meaning of relatively straight forward parameters. The following list is intended to give a basic understanding of what parameters are stated by Murata Power Solutions.

## 2. PARAMETERS

### INDUCTANCE (L)

The classical definition of inductance is a constant that relates the magnetic flux linking a circuit to the current flowing in the circuit. The inductance is measured as a reactance to an AC signal as a single frequency (typically 10mV at 1kHz).

### MAXIMUM DC CURRENT ($I_{DC}$)

The maximum DC current is defined as the DC current at which the inductance falls to 90% of its nominal value (see Figure 1). However, this does not recognise the effect of self-heating also contributing to a change in inductance (usually inductance falls as the temperature rises). Consequently, the maximum DC current is limited to a temperature rise of 30°C. Therefore, the maximum DC current is the value at which the inductance falls to 90% of its nominal value or until the temperature rise reaches 30°C, whichever is sooner. When making comparisons with inductors it is worth noting that many manufacturers quote a 30% drop which gives the impression of a higher current rating. Murata Power Solutions inductors will withstand current spikes greater than $I_{DC}$ for short periods

### SELF RESONANT FREQUENCY ($f_0$)

Self resonance occurs when the impedance of the inductor is purely resistive. At this frequency the capacitive effect of the wire and the inductance cancel and the relative signal phase across the inductor is zero

### QUALITY FACTOR (Q)

In single reactive components the quality factor is usually the ratio of the reactance and resistance (ideally $Q= \omega L/R_{DC}$, where $\omega=2\pi f$). The value quoted in the specification is a measured value at a specific frequency this compensates for the capacitance of the wire (all measurements are made on a Wayne-Kerr WK3260B impedance analyser with a 10mV signal).

### INDUCTANCE TEMPERATURE COEFFICIENT

The change in inductance per unit temperature change. Measured under zero bias conditions and expressed in parts per million (ppm).

### RESISTANCE TEMPERATURE COEFFICIENT

The change on DC wire resistance per unit temperature change. Measured under low DC bias (<1VDC) and expressed in parts per million (ppm).

### CURRIE TEMPERATURE ($T_C$)

The temperature beyond which the core material loses its magnetic properties.

### MAGNETIC SATURATION FLUX DENSITY ($B_{SAT}$)

A core parameter which isolates the maximum flux the material con be induced to hold. At this value of flux density all magnetic domains with the core are magnetised and aligned.

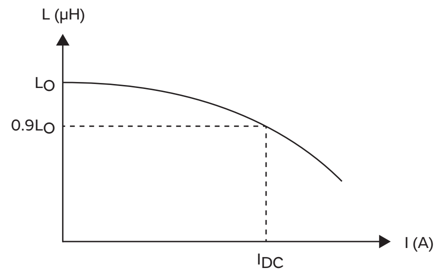


FIGURE 1. MAXIMUM DC CURRENT

## 3. OTHER PARAMETERS

### SATURATION

Saturation of an inductor occurs when the core can no longer store magnetic energy, (energy storage = ½ $LI^2$).

### EMI

Electromagnetic interference in inductors refers to the amount of magnetic field radiated away from the inductor itself, that is into "space". This field may cause interference with other magnetically sensitive components and requires considerations in circuit design and layout and may determine the selection of inductive components in certain applications.

## 4. INDUCTOR TYPES

### BOBBIN

The bobbin inductor (shaped like a cotton bobbin) has open sides, hence a large air-gap in the magnetic circuit; from top-to-bottom of the bobbin edges. This shape supports high currents at high inductance and very low losses. Open bobbin shapes can exhibit higher magnetic EMI around their open sides in their air-gap but are usually the lowest cost form due to ease of machine winding.

The EMI performance of bobbin inductors can be improved by adding a magnetic shield around the bobbin, hence the air-gaps are only at the top and bottom of the shield. The shield increases the inductance per turn due to more magnetic material, but reduces the saturation current compared to a similar sized open-frame bobbin type.

### TOROID

Toroidal inductors exhibit very low EMI, the shape of the core means the magnetic path flows only with in the inductor and hence stray field is virtually eliminated. At high inductance values saturation can occur at relatively low current.
The most common form used in switching regulators due to their EMI properties, toroid inductors are also useful in applications where saturation can be used as either a limiting or feedback mechanism (e.g. chokes, power filters). Toroids can be used at up to several MHz with careful choice of core material.

### POT CORES AND OTHER FORMS

Pot cores form an enclosed shape around the winding and typically come in 2-parts. Most of the flat-coil inductors from Murata feature a form of pot-core (named after their "cooking pot" shape). The shape forms a very low EMI magnetic circuit as the magnetic field is fully enclosed within the core with small air-gaps only at the join between the 2-halves of the pot-core.

The description for pot-cores can be applied to forms that separate from the original format, such as flat E-I cores, hence may be generically applied to other shapes that have similar confined magnetic fields with small air-gaps at the joins of the magnetic materials.

## 5. ADVANTAGES OF PASSIVE NETWORKS

It is easy in the "silicon age" to dismiss inductors as circuit elements in favour of what are considered cheaper, physically smaller and lighter active networks. However, there are many properties of inductors which cannot be produced using "cheap" silicon, one that is immediately brought to mind is use in power circuitry. Another area in which an inductor may prove cheaper is in a simple filter circuit, filters are usually considerably more complex and often relatively expensive to implement as simple low order filters compared to passive inductive designs.

At high frequencies, silicon and RC networks become limited, stray capacitance and transistor switching frequencies can restrict design capability and increase cost, amplifier stability can also cause serious problems and grounding configurations require special attention. While not alleviating all the problems associated with high frequency design, low DC resistance of inductors are easily characterised and predictable frequency responses can make circuit design and analysis of circuit behaviour easier.

There are certainly some definite advantages in using inductors in modern circuit designs and these passive elements should not be neglected. The following design notes give some basic ideas for use of the Murata range of inductors. The designs can be used as shown or alternatively the basic ideas adapted to the readers own requirements. The notes are not exhaustive and we welcome feedback and application ideas from any customers, similarly we are always willing to discuss your requirements and can, if required, custom design an inductor for your application.

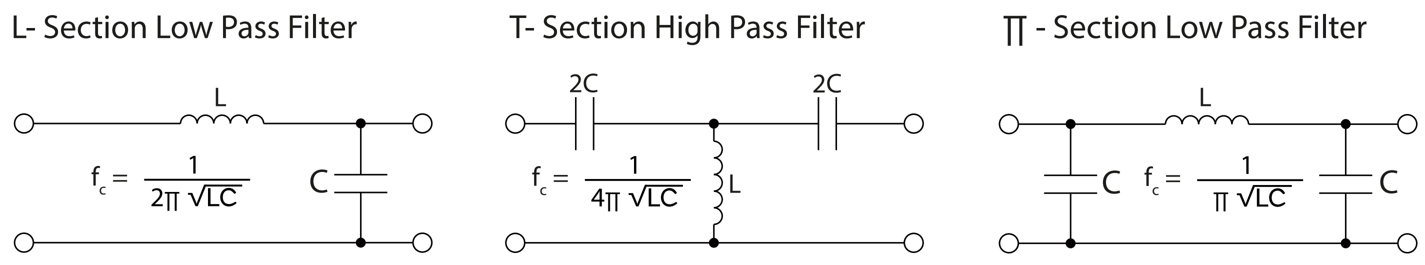
## 6. INDUCTORS IN PASSIVE FILTERS

### SIMPLE FILTERS

Simple filters can be easily constructed with inductor and capacitor elements. These filters have the advantage of being easy to calculate and characterise the frequency response. They exhibit few of the secondary effects and stability problems associated with their active counterparts. In power line filtering, inductors and relatively small capacitors can be used to produce a smoothed ripple from a spiked input response, ideal for recurring noise in switching regulator circuits and power lines in noisy environments.

Simple L, T and $\pi$ section filters can be constructed to provide low, high and band pass functions (see Figure 2). There are many texts on this subject and these should be consulted for details on more complex filters and interaction between sections. These pages are simply illustrative of applications Murata Power Solutions inductor parts.

**Figure 2. SIMPLE FILTER CONSTRUCTION**



## 7. INDUCTORS IN SWITCHED MODE POWER SUPPLIES (SMPS)

There are many silicon switching circuits available on the market from most analogue silicon vendors (Texas Instruments, Analog Devices Inc.). All the design notes which accompany these devices recognise the importance of correct inductor selection to achieve the optimum performance of the SMPS design.

## 8. BASIC TOPOGRAPHY

Essentially there are 2 types of switching converter; a boost converter switching from a lower to a higher voltage, and a step-down (buck) switching from lower to higher voltage. There are also buck-boost which can do both, useful for when the input may dip for short periods, but for our purposes we will here only look at the buck and boost as separate converters as these are the most common applications.

There are multiple controller ICs that can provide all the control and switching required that are available from multiple analogue IC vendors. Which to use will be up to the application and requirements for other features (e.g. short-circuit protection, over-voltage protection), but essentially they all have a similar topology; there is a switch for the power through the inductor, an oscillator which determines the frequency and pulse shaping and a control algorithm that determines how the oscillator will behave as the output voltage reaches the required target voltage (see Figures 3 and 4).

Different ICs tend to have different control methods, although the most common are relatively straight-forward.

### PULSE SKIPPING

This is where a constant frequency and pulse width is used, but as the output nears the target, or the load is light, the controller misses out pulses (skips) to maintain regulation of the output voltage.

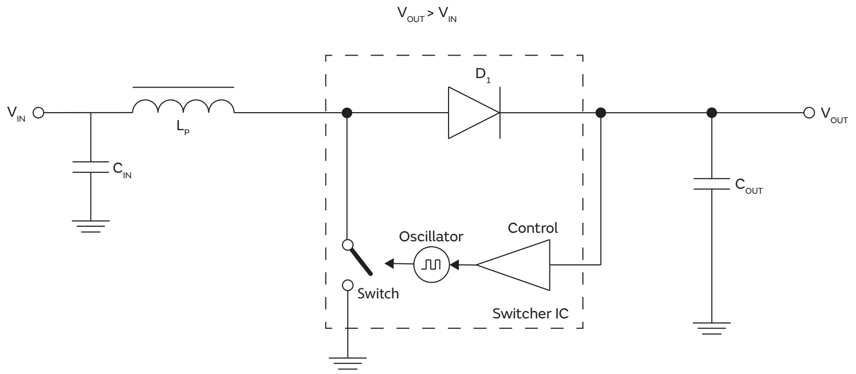
### PULSE WITH MODULATION

In this method the frequency is usually kept constant, but the width (time) of the switched pulse is narrowed to be very short as the output nears the target voltage or as the load reduces.

### SPREAD SPECTRUM FREQUENCY MODULATION

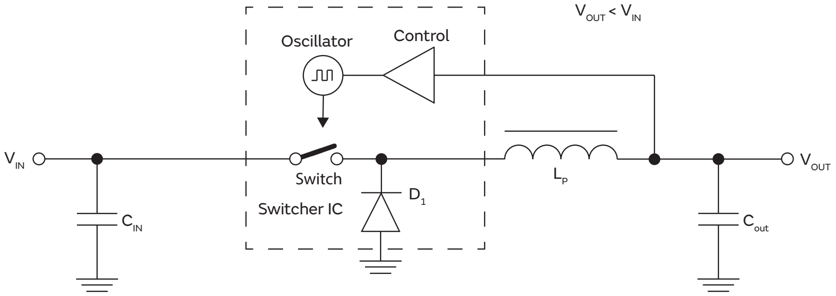
This is often applied on-top of one of the above control strategies and is used to reduce the EMI of the circuit. It essentially "jitters" the frequency around the selected switching frequency, typically by a few percent. This spreads out the switching noise in the frequency domain, but be aware this is only effective if average or quasi-peak mode is allowed in your EMI test regime, if using peak-detect this methodology does not improve the EMI at testing.

**Figure 3. BOOST CONVERTER TOPOLOGY**



Note: sometimes the diode ($D_1$) may be external to the IC and/or may be replaced with a transistor switch; known as synchronous rectification

**Figure 4. BUCK CONVERTER TOPOLOGY**



Note: sometimes the diode ($D_1$) may be external to the IC and/or may be replaced with a transistor switch; known as synchronous rectification

## 9. INDUCTOR SELECTION

The inductance value for a switching regulator must be high enough to prevent excessive current through the diode and low enough to store sufficient energy in the core. The DC resistance should be low to reduce losses and prevent self heating. The core must also be capable of storing the required energy without saturating. For general purpose use bobbin wound inductors are excellent for most cases, however, in EMI sensitive applications a toroid or sheilded bobbin may be preferable.

The peak current ($I_{PK}$) and inductance ($L_P$) value should be calculated for two worse case conditions (i.e. maximum and minimum values of $I_{PK}$ and $L_P$). The final choice should be an inductor whose inductance is in between the limits of the calculated values and which has a DC current rating in excess of the maximum peak current calculated.

(1) $$I_{PK} = \frac{4\ I_{OUT}\ (V_{OUT}-V_D)}{(V_{IN}-V_{SW}-V_D)}$$

(2) $$L_P = \frac{t_{ON}\ (V_{IN}-V_{SW}-V_{OUT})}{I_{PK}}$$

Where $V_{OUT}$ is the required output voltage, $V_{IN}$ is the input supply voltage, $I_{OUT}$ is the required output current (load current), $V_{SW}$ is the voltage drop across the switching element and $V_D$ is the voltage drop in the diode.

## 10. DESIGN EXAMPLE

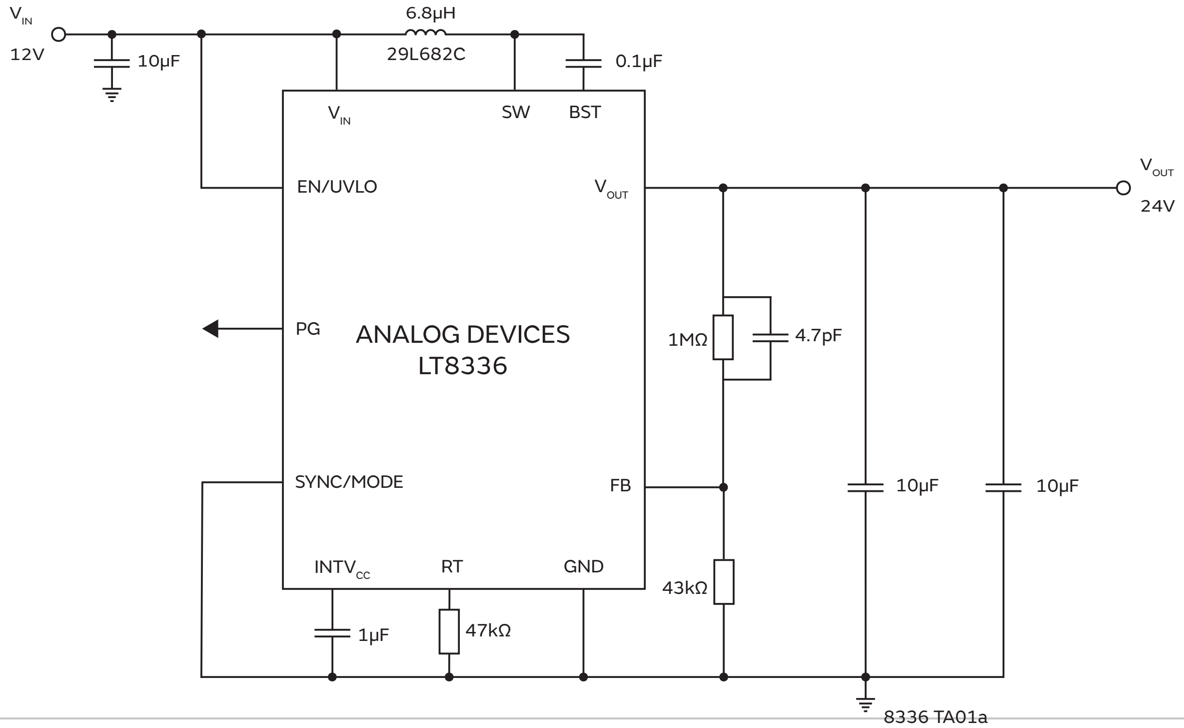
### DESIGN EXAMPLE 1: BOOST CONVERTER

Figure 5 shows a boost converter, taking a 12V input and providing a 24V output, typically an industrial processing or measurement circuit. In this application the frequency of the switching is determined by the resistor connected to the RT pin of the Analog Devices LT8336 IC, using a 47kΩ value gives a switching frequency just above 2MHz, and avoids the AM radio band. In this configuration with the 29L682C inductor the circuit can in theory deliver around 30W of power, however, this is limited by the input voltage stability and its regulation range and under 20W is more realistic power delivery for this circuit.

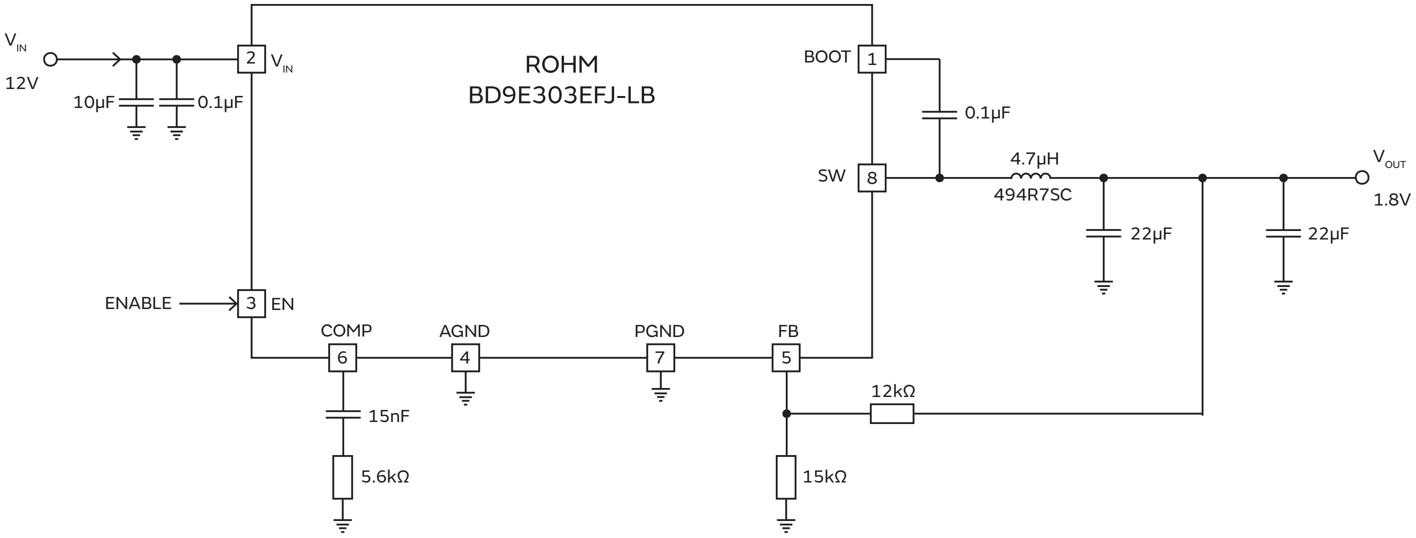
### DESIGN EXAMPLE 2: BUCK CONVERTER

Figure 6 shows a buck converter, taking a 12V input and providing a 1.8V output, this would typically be direct to a micro-processor or digital signal processor. In this application the frequency of the switching is set at approximately 300kHz by the IC; Rohm BD9303EFJ. The resistor-capacitor combination connected to the COMP pin sets the phase compensation but does not impact the frequency directly. In this configuration with the 494R7SC inductor the circuit can deliver around 5W of power, limited by the switch current capability of the IC.

## FIGURE 5. DESIGN EXAMPLE BOOST CONVERTER



## FIGURE 6. DESIGN EXAMPLE BUCK CONVERTER



## 11. MODELLING INDUCTORS IN SPICE

If a SPICE simulation of an inductor is required the simple standard value may not be adequate for all occasions, particularly in situations where the inductor is being used either near its maximum DC current value or when operated at near resonant frequency. In a circuit using many low resistance parts or where low resistance is significant the effect of the DC resistance of the inductor may be noticeable in the circuit, especially when using high value inductors that have high DC resistance. The specification for an inductor gives sufficient date to accurately model some of the more complex effects of an inductor in a circuit.

An inductor model can be constructed from 4 passive elements, an ideal series inductor and resistor, a parallel capacitor and a parallel resistor (see Figure 5). The value of the series resistor is defined by the DC resistance value in the device specification (e.g. for a 1900 series 100µH inductor, $R_{DC} = 0.065\Omega$). The value of the parallel capacitor can be derived from the self resonant frequency ($f_0$) of the part, since this is the point at which the reactance of the inductor is zero (i.e. the impedance is purely resistive). At self resonance the capacitance is given by;

$$(3)\ C_P = \frac{1}{(2\Pi f_0)^2 L_0}$$

Magnetic core loss is modelled as a parallel resistor ($R_P$) across the terminals. The value can be calculated from the quality factor (Q);

$$(4)\ R_P = Q(2\Pi f_0)^2 L_0$$

This parallel resistor limits the simulated self resonance rising to ‘infinity’. The model with these four basic circuit elements models the inductors impedance and phase behaviour over a wide frequency range (see Figures 7 and 8).

The inductance value is not constant as the SDC current through the device reduces its inductance, since part of the core is magnetised (Figure 1). The standard inductor model in version 2G6 of SPICE can accommodate this effect as a polynomial expression for the inductance as a function of current (equation 5), this can be entered in the standard inductor description in SPICE.

$$(5)\ L_I = L_0 + L_1 I + L_2 I^2 + \dots + L_n I^n$$

Where $n \leq 20$.

A simple 2nd order polynomial is sufficient to model this effect and the maximum DC current value can be used to determine the coefficient L2. If the inductance is 90% of its nominal value, $L_0$, at the maximum DC current IDC, then the polynomial equation is;

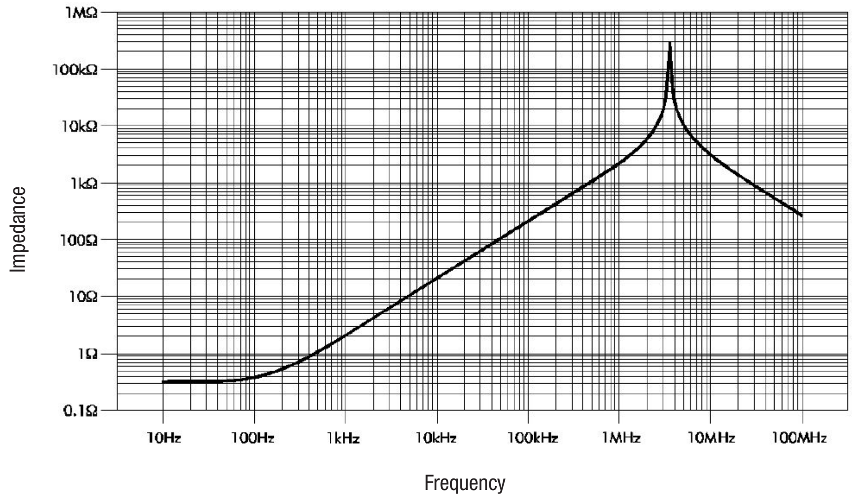
$$(6)\ 0.9 L_0 = L_0 + L_2 I_{DC}^2$$

Hence the coefficient is given by;

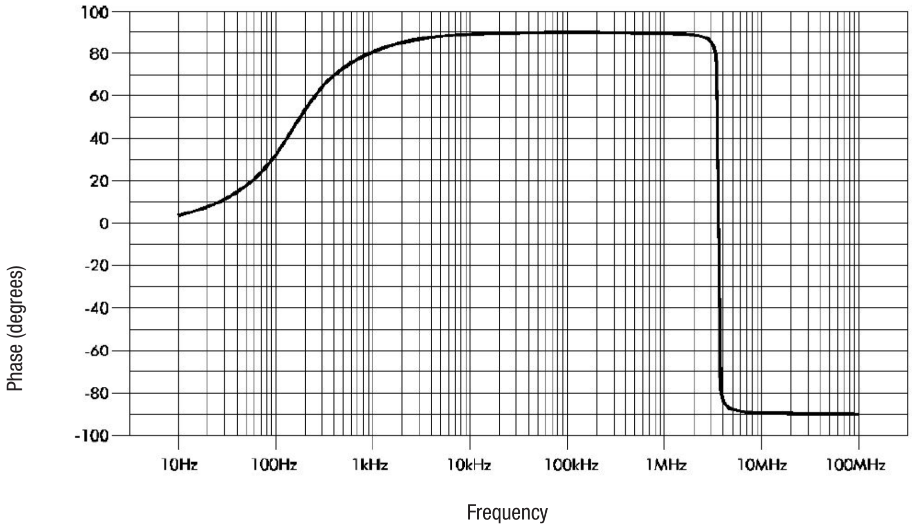
$$(7)\ L_2 = -\frac{0.1 L_0}{I_{DC}^2}$$

An accurate and relatively complex model for an inductor can now be constructed using datasheet values only, hence no additional measurements by the user are required.

## FIGURE 7. 18R334 IMPEDANCE SIMULATION



## FIGURE 8. 18R334 IMPEDANCE SIMULATION



## 12. SPICE MODEL EXAMPLE

SPICE EXAMPLE SUB CIRCUIT NETLIST

The following example is a model for a 1800R series 330µH inductor (18R334) where $L_0$ = 330µH, $R_{DC}$ = 0.315Ω, $I_{DC}$ = 0.73A, Q = 40 and $f_0$ = 3.57MHz.

```
.SUB CKT L18R334 1 2
LO    3 2 POLY 330E – 6 0 – 6.19E – 5
RDC  1 3 0.315
CP    1 2 6.02E – 12
RP    1 2 363K
.ENDS L18R334
```

## 13. LIMITATIONS OF SPICE MODEL

The above model is now quite sophisticated for an inductive element, however, there are still limitations and this should be borne in mind. The model assumes that there is no variance of resistance and capacitance with DC current, at low values of these parameters this may be adequate as these will tend to be swamped by the rest of the circuit. The major limitation is in the lack of temperature modelling of the inductor, however, this is a general limitation of SPICE. Some temperature modelling for the resistor could be incorporated, however, the heating effect of the power dissipated in the inductor ($I^2R$ term) is not modelled.

It is possible to simulate several complex aspects of inductor operation using only 3 additional passive elements and a simple polynomial expression. The resulting model givers accurate inductor simulations in SPICE over a wide range of operating conditions with a minimal increase in computation time (only one extra node is introduced).

## 14. EMC DESIGN CONSIDERATIONS

The regulations regarding electromagnetic compatibility (EMC) will affect many aspects of circuit and system design.

However, there are many considerations that can be applied generally to reduce the emissions from and susceptibility to electromagnetic interference (EMI).

There are many areas where the use of inductors in decoupling and filtering applications will help. However, there may also be situations where the inductive element in a switching circuit is the major cause of EMI. As a manufacturer Murata is committed to minimising emissions from its own components and to helping its customers achieve EMC compatibility by correct component choice and design, to this end we have compiled the following list of general design considerations.

- Reduce high frequency (particularly radio frequency) loops in supply lines.
- De-couple supply lines at local boundaries (use RCL filers with $Q \leq 2$).
- Use low pass filters on signal lines to reduce band width to signal minimum.
- Keep return and feed loops close on wide bandwidth signal lines.
- Terminate lines carrying HF or RF signals correctly (this minimises reflection, ringing and overshoot).
- Avoid slit apertures in pcb layout, particularly in ground planes or near high current paths.
- Use common mode chokes between current carrying and signal lines to increase coupling and cancel stray fields.
- Use discrete component and filters where possible.
- Ensure filtering of cables and over voltage protection (this is especially true of cabling that is external to the system, if possible all external cabling should be isolated and terminated at the equipment boundary).
- Isolate individual systems where possible, especially analogue and digital systems, on both power supply and signal lines.
- If available, use shielding on fast switching circuits, mains power supply components and low power circuitry.

In general, keeping the bandwidth of all parts of the system to a minimum and isolating circuits where possible reduces susceptibility and emissions.