AC 2009-352: THREE PRACTICAL AND EFFECTIVE RF AND EMC EXPERIMENTS FOR A COMPUTER ENGINEERING COURSE ON ELECTROMAGNETICS AND EMC

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Three Practical and Effective RF and EMC Experiments for a Computer Engineering Course on Electromagnetics and EMC

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Abstract

This paper presents three practical and effective electronic hardware experiments which demonstrate respectively (1) use of a common-mode choke to perform common-mode current suppression, (2) construction and analysis of an FM wireless microphone and (3) study of bypass capacitor effectiveness to demonstrate concepts related to electromagnetic compatibility (EMC), signal integrity (SI) and radio frequency (RF) design. The experiments have been performed as part of the laboratory portion of our required junior-level course for computer engineering students on electromagnetics and EMC. These experiments have helped students understand the underlying physics in addition to demonstrating measurement techniques and solution options for the topics discussed in the course. They are easy and inexpensive to implement and perform because they can be set up on a standard breadboard. They should prove useful in any engineering course on RF circuits, electromagnetic, SI and EMC. As high speed, low power, wireless, and hand-held embedded engineering designs become more common, computer engineering students have a growing need for knowledge and experience in design and manufacturing issues related to SI and EMC. We have found these experiments to be valuable and effective in enhancing student interest in RF, SI and EMC.

1 Introduction

Electromagnetic compatibility (EMC) and signal integrity (SI) have become pervasive design issues in high-speed designs of digital systems, wireless devices, mixed signal systems, and hand-held devices. Two ready examples of industries with acute and long-standing need for engineers with an understanding of EMC design issues are automotive electronic systems and home entertainment systems.

Ensuring signal and power integrity in digital electronic systems is vital to maintaining signal viability as these signals propagate along their path amidst noise from different sources. The topic is critical in electronic system design at various levels including backplanes, printed circuit boards, integrated circuit packages and integrated circuits. Signal and power integrity are inherently multidisciplinary topics that draw upon knowledge and techniques from circuit design, electromagnetic field theory, material properties, and packaging design.

As signal speeds increase, lumped element models become inadequate to describe circuit behavior, and coupling and crosstalk between adjacent conductors become significant problems. The models used to describe electronic signals and devices must consider these electromagnetic effects at high speeds when the sizes of devices and signal paths are of the same order as wavelength of the signals. Circuit designers will need to understand that signals propagate as waves in order to successfully design high speed circuits and systems. Signal integrity has been taught at universities primarily as senior elective or graduate level classes¹⁻⁵. A Master degree program in signal integrity has been established at University of South Carolina in collaboration with Intel⁶. There is an urgent need to develop teaching and evaluating materials for signal integrity and EMC for engineering courses, especially at undergraduate level. The EMC Society of the IEEE has published an EMC laboratory manual⁷. This paper is our effort to enhance the experimental material for teaching EMC and SI with three practical and effective electronic hardware experiments to demonstrate EMC and RF effects: the first on common-mode current suppression, the second on an FM wireless microphone, and the third on bypass capacitor effectiveness. Each experimental circuit may be constructed on a standard wireless circuit prototyping breadboard.

The first experiment demonstrates how to design and evaluate a common-mode choke that is characterized in terms of self and mutual inductances L and M. The choke is installed on a switching-mode DC-DC converter built from discrete components. The conversion efficiency of the converter is measured at different switching frequencies. Common-mode currents from the converter are measured with a current probe clipped on the DC power cable both with and without the common-mode choke. Also, conducted emissions on the 120 VAC power line are measured with a line impedance stabilization network (LISN). Clear benefits of using the common-mode choke are demonstrated using both the LISN and the current probe.

The second experiment analyzes an FM wireless microphone circuit operating at 102 MHz which is generated as the third harmonic of a 34 MHz signal from a RF oscillator. This design is easily built in the laboratory by the students, since it can be built on a standard prototyping breadboard at 34 MHz. A audio amplifier input slightly changes the bias point on the RF oscillator transistor at an audio rate, which allows the audio signal to affect frequency modulation by changing the transistor junction capacitance, which in turn varies the oscillation frequency at an audio rate. Once built, this experimental platform can be used to evaluate radio wave propagation, antenna polarizations and antenna lengths.

The third experiment shows how bypass capacitors on a DC power bus affect power bus noise voltage and also radiated emission from the bus. One clear conclusion from this experiment is that capacitors with larger capacitance do not necessary produce better decoupling results. The students begin by building a 3-inverter ring oscillator using a 74HC00 digital hex inverter IC. The oscillator is powered by a 5 VDC power supply with a 0.1 μ F bypass capacitor placed as close as possible across its Vcc and Ground power pins. Noise voltage on the power supply is observed on an oscilloscope and radiated emissions up to 600 MHz are observed on a spectrum analyzer as different bypass capacitor combinations are placed across the power and ground pins. Students observe that capacitors with larger values can be less effective than those of smaller values because of larger inductive effects and lower resonant frequencies of the larger capacitors.

These experiments work together to expose students to many common topics in EMC such as common-mode currents, switching power supply design, conducted emission, antenna length, FM modulation, harmonics of a rectangular waveform, mutual and self inductances, audio amplifier, bypass capacitors, and resonant frequency of a capacitor. In addition, many connections are made to other prerequisite classes such as circuits, electronics, and system analysis.

The experiments have been used as both demonstrations and student laboratory exercises in our required junior-level course on electromagnetics and electromagnetic compatibility (EMC) for computer engineering students. They have helped to enhance student's understanding of the underlying physics, problem characteristics, solution options, and measurement techniques for the topics under discussion. They are also easy and inexpensive to implement and perform. They should prove useful in other engineering courses on RF circuits, electromagnetics and EMC.

2 Experiment #1: A Common-mode Choke in DC-DC Converter Design

The first experiment demonstrates how to build and evaluate a common-mode choke that is characterized in terms of self and mutual inductances L and M and is installed between a DC power supply and a switching DC-DC converter built from discrete components. The conversion efficiency of the converter is measured at different switching frequencies. Common-mode currents from the converter are measured with a current probe on the DC power cable both with and without the common-mode choke. Finally, conducted emissions on the 120 VAC power line are measured with a line impedance stabilization network (LISN). Clear benefits of using the common-mode choke are demonstrated using both the LISN and the current probe.

The objectives for this experiment and instruments used are described in Section 2.1. The construction and measurement of the choke is presented in Section 2.2. The construction and measurement of the DC-DC converter is shown in Section 2.3.

2.1 Objectives and Equipment

The main objectives of this lab are listed in Table 1. They are composed of four topics: mutual inductance, common-mode choke, switching-mode DC-DC voltage conversion, and conducted emission and its mitigation. In the lab, students integrate knowledge gained from previous prerequisite classes as they develop an understanding of the vital topics of common-mode current and conducted emissions.

Table 2 lists the instruments needed to perform this lab. Most of them are common lab instruments that can be found in an electronics lab of an engineering college. The only appliances that might not be common are the Agilent 3 GHz spectrum analyzer and the EG&G 500 MHz snap-on current probe.

2.2 The Construction and Measurement of the Common-Mode Choke

The common-mode choke was constructed by bifilar winding 20 turns of 2 strands of 20-gauge hookup wire around a ferrite toroidal core as shown in Figure 1. The ferrite core used in this experiment has the following specifications: outer diameter of 2.5 cm, inner diameter of 1.0 cm thickness of 0.9 cm and relative permeability μ_r of 5000. Either the Amidon FT-114-J or the FT-140A-J is a suitable replacement, though the core dimensions will change. (Amidon Inc., P.O. Box 6122, Irvine CA 92616, Phone 714 850 4660, www.amidoncorp.com.)

Table 1.	Objectives of	the Common Mo	de Choke Experiment
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Measure self-inductance and compare it with a predicted value.
Understand operation of common-mode choke.
Measure self-inductance L and mutual inductance M of a common-mode choke.
Analyze and construct a switching-mode DC-DC voltage converter.
Measure the conversion efficiency of the converter at different switching rates.
Verify effectiveness of common-mode choke on common-mode current of the converter.
Observe effectiveness of the common-mode choke on conducted emission from the
converter.

Table 2.	Instruments	Used in the	Common Mode	Choke Experiment
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Agilent E4402B ESA-E Series 100 Hz – 3 GHz Spectrum Analyzer
EMCO Model 3810/2 LISN (9 kHz – 30 MHz)
Agilent 54624A 100 MHz Digital Oscilloscope
EG&G SCP-5(I)HF (125 kHz – 500 MHz) Snap On Current Probe
Agilent E3631A Triple Output DC Power Supply (5 V at 5 A)
Agilent 33250A 80 MHz Function Generator
Assorted connecting cables and adaptors
An Extech Model 380193 LCR meter

The self inductance L of the toroidal coil is obtained in three ways: by a LCR meter, by putting it in series with a known capacitance and measuring the series resonant frequency, and by a formula. All three estimations produced numbers that are within 10% of each other. The LCR meter Extech Model 380193 measured the self inductance as 3.51 mH. The resonant frequency measurement of the single coil in series with a 10 nF capacitor using a function generator and a scope yielded a self inductance of 3.48 mH. The self inductance is found to be 3.3 mH using the following formula:



Figure 1. The Common-mode Choke Construction.

Another method of determining both self and mutual inductances simultaneously involves the measurement of the resonant frequencies of the two configurations of the coil in series with a 10 nF capacitor as shown in Figure 2. From the measured equivalent inductances $L_{eq}(AC)=2(L-M)$ and $L_{eq}(AD)=2(L+M)$, L and M may be calculated. The measured values of $L_{eq}(AC)$ and $L_{eq}(AD)$ were 0.59 µH and 15 mH respectively. The calculated values for L and M were L=3.7502 mH and M = 3.7499 mH. Note that L and M are almost the same, indicating an excellent common-mode choke.

The choke exhibits a high inductance of Leq = 2*(L+M) = 15 mH for common-mode current which needs to be reduced and a low equivalent inductance of Leq = 2*(L-M) = 600 nH for differential mode current which needs to be transmitted. Thus differential-mode signal currents would pass through the choke easily, and common-mode noise currents would be blocked by the choke.

2.3 The Construction and Measurement of the DC-DC Converter

A switching-mode DC-to-DC voltage converter was constructed and tested to show the effectiveness of the common-mode choke in reducing conducted emission from the converter. The schematic diagram of the converter is shown in Figure 3. The converter is composed of an astable blocking oscillator, a toroidal CT transformer and a half-wave rectifier with a capacitor filter.

When the DC power for the astable blocking oscillator is turned on, both BJTs turn on equally, as base-emitter current flows through R1 and R2. However, this is a potentially unstable circuit (like a ball resting on the crest of a hill). Imagine that there is a small positive noise glitch at the base of, say, Q2, that suddenly makes the base current of Q2 slightly greater than that of Q1. This will cause Q2 to conduct more than it was, lowering the voltage at the collector of Q2 and (since the voltage across C2 cannot change instantly), lowering the base voltage on Q1, causing Q1 to conduct less than it was. This raises the voltage at the collector of Q1, and this in









Figure 2. Measuring L and M of the Common-Mode Choke

turn makes Q2 conduct even harder. This positive feedback situation quickly drives Q2 in saturation and Q1 into cutoff. However Q2 does not remain saturated and Q1 does not remain cut off for long. This is because C2 charges through R2, and when C2 is charged high enough so that V_{b1} exceeds Q1's cut-in voltage, $(V_{be})_{CUT-IN}$, then Q1 turns on, and this causes Q2 to turn off. The process then reverses, with C1 charging until Vb2 exceeds Q2's cut-in voltage, etc.

Therefore continuous oscillation occurs, with Q1 and Q2 alternately changing between saturation and cut off. While Q2 is saturated during the first half of the oscillation period, current first flows from the center tap to the right terminal of the toroidal transformer, and then when Q1 saturates during the second half of the period, current flows from the center tap to the left terminal of the transformer, allowing a higher voltage to be induced in the secondary coil by transformer action. To find the time it takes for V_{b1} to increase from its initial value to its cut in value, we must think back to the point in time just before Q2 saturates. At this time we may assume that Q1 is saturated and Q2 is cut off. Then at this time the voltage on the right side of C2 is Vin, and the voltage on the left side of C2 is $(V_{be})_{SAT}$. Just before Q2 saturates, the voltage across capacitor C2 is given by $(V_{b1} - V_{c2})_{init} = (V_{be})_{SAT} - Vin$. This must also be the voltage across capacitor C2 just after Q2 saturates, since capacitor voltage cannot change instantly.

Therefore, at the point in time just after Q2 saturates, the value of V_{b1} is given by

$$(V_{b1})_{init} = (V_{b1} - V_{c2})_{init} + (V_{ce})_{SAT} = (V_{be})_{SAT} - Vin + (V_{ce})_{SAT}$$

 V_{b1} charges toward a value of $(V_{b1})_{final}$ = Vin, though it never gets there, since when V_{b1} = $(V_{be})_{CUT-IN}$, the states of the transistors change. Because R2 = R1 = R and C2 = C1 = C, and the BJTs (Q1 and Q2) are matched, it takes $\frac{1}{2}$ period = T/2 for V_{b1} to charge from $(V_{b1})_{init}$ up to $(V_{be})_{CUT-IN}$.

Using the general first-order RC switching transient formula:

$$Vx(t) = Vx_{final} - (Vx_{final} - Vx_{init}) \cdot exp(-t/RC))$$

we see that $V_{b1}(t) = Vin - (Vin - [(V_{be})_{SAT} - Vin + (V_{ce})_{SAT}]) \cdot exp(-t/RC))$

When $\frac{1}{2}$ period elapses (t = T/2), $V_{b1}(t) = (V_{be})_{CUT-IN}$, so

$$(V_{be})_{CUT-IN} = Vin - (Vin - [(V_{be})_{SAT} - Vin + (V_{ce})_{SAT}]) \cdot exp(-(T/2)/RC))$$

Solving for the period of oscillation, T

$$\mathbf{T} = -2 \cdot \ln[\frac{\mathrm{Vin} - (\mathrm{V_{be}})_{\mathrm{CUT-IN}}}{2 \cdot \mathrm{Vin} - (\mathrm{V_{be}})_{\mathrm{SAT}} - (\mathrm{V_{ce}})_{\mathrm{SAT}}}] \cdot \mathbf{R} \cdot \mathbf{C}$$



Figure 3. A Switching-mode DC-to-DC Converter with Common-mode Choke.

Since Vin >> $(V_{be})_{CUT-IN}$, $(V_{be})_{SAT}$, and $(V_{ce})_{SAT}$, we may make an approximation to the above expression for the period of oscillation:

$$\mathbf{T} = -2 \cdot \ln[\frac{1}{2}] \cdot \mathbf{R} \cdot \mathbf{C} = 2 \cdot \ln(2) \cdot \mathbf{R} \cdot \mathbf{C}$$

The approximate frequency of oscillation is therefore

$$\operatorname{fosc} = \frac{1}{\mathrm{T}} = \frac{1}{2 \cdot \ln(2) \cdot \mathrm{R} \cdot \mathrm{C}}$$

Here $R2 = 15 \text{ k}\Omega$, and $C2 = 0.01 \mu\text{F}$, therefore

$$\text{fosc} = \frac{1}{2 \cdot \ln(2) \cdot 15000 \cdot 0.01 \times 10^{-6}} = 4.809 \,\text{kHz}$$

The measured value of fosc was 11 kHz, so the approximate result obtained from the above analysis is not particularly accurate. This is probably due to the fact that the above derivation did not take into account the effects of the inductive load (the R-L time constant), which would alter the initial value of V_{b1} that was assumed.

The conversion efficiency is defined as the ratio of the load power to the input power, $Pload/Pin = (Vin*Iin) / ((Vload)_{AVG}^2 / Rload)$. The conversion efficiency of the converter at different switching frequencies is shown in Table 3.

C2	fosc	Pin	Pout	Efficiency
0.047 μF	4.33 kHz	1.85 W	1.04 W	56.1%
0.01 µF	11.6 kHz	2.14 W	1.31 W	61.1%
0.001 µF	29.1 kHz	3.19 W	1.19 W	37.1%

Table 3. DC-to-DC Conversion Efficiency at Different Switching Frequencies. (C2 is changed while R2 remains at $15 \text{ k}\Omega$)

Two breadboard circuits of the converters with and without the choke are shown in Figures 4 and 5, together with their corresponding common-mode current measurements. An EG&G SCP-5(I)HF snap-on current probe with frequency range from 125 kHz to 500 MHz was clamped around the power supply wires to measure conducted radiation on them, as shown in Fig. 4. The common-mode choke was found to reduce the magnitude of the common-mode currents by 5 to 10 dBm over frequency range from 9 kHz to 50 MHz. Conducted emissions of both circuits were measured with a line impedance stabilization network (LISN), as shown in Figure 6. There is appreciably less conducted emission when the choke was inserted into the circuit.



Figure 4. Common-mode Current of the Circuit without Common-mode choke.



Figure 5. Common-mode Current of the Circuit with Common-mode Choke.



Figure 6. 60 Hz Power Line Conducted Emissions without and with the Commonmode Choke Measured with LISN.

3 Experiment #2: An FM Wireless Microphone

This experiment is to build and evaluate an FM "wireless microphone" transmitter operating at first 34 MHz and then at 102 MHz.

The schematic diagram of an RF oscillator operating at 34 MHz is shown in Figure 7. The 102 MHz third harmonic of the 34 MHz oscillator circuit is transmitted and received by a conventional FM radio receiver. The initial circuit was designed to oscillate at 34 MHz, because it was feared that building an RF oscillator circuit on a conventional circuit prototyping breadboard would not permit oscillation directly at 102 MHz. This is because of the appreciable inductance of the individual breadboard traces and the sizeable capacitance (about 5 pF) between adjacent traces. (This fear turned out to be largely unfounded, as long as the circuit was laid out and wired with very short wires, as will be discussed later.) Modulation is achieved by capacitively coupling the output of the audio amplifier to the base of the RF oscillator transistor, which varies the bias point of the RF oscillator thereby achieving FM modulation by varying the junction capacitance of the oscillator transistor at an audio rate.



Figure 7. Complete Circuit of FM Wireless Transmitter.

wireless microphone circuit was about 50 feet, though it varied with the transmitting antenna length. This experimental platform was then used to evaluate radio wave propagation with different antenna polarizations and different antenna lengths.

In this lab, students design and measure a $1.0 \,\mu\text{H}$ solenoidal air-core inductor, build an audio microphone amplifier circuit, build a typical RF "LC" oscillator circuit, and add an audio modulation circuit to complete a wireless microphone. They would then measure harmonic suppression and evaluate radiation spectrum of the transmitter.

3.1 Objectives and Equipment

The main objectives of this lab as listed in Table 4 are to gain insight into inductor, audio amplifier, FM modulation, antennas and radiation patterns.

Table 5 lists the instruments needed to perform this lab. The only uncommon instrument is the curve tracer in addition to the spectrum analyzer.

3.2 Construction and Measurement of a 1µH Solenoidal Air-core Inductor

A 1.0 μ H air-core solenoidal-wound and single-layer inductor is designed and constructed with a suitable coil form such as a felt-tip marker pen and insulated hookup wire. Its inductance is estimated by the following formula: $L = \frac{N^2 \mu A}{1}$, where N = Number of turns, A = cross-sectional area of coil, and 1 = length of coil in air $\mu = \mu_0 = 4\pi \times 10^{-7}$ H/m. An inductor of about 1 μ H was made with a coil form of 0.8 inch in diameter, 1.2 inches in length, 9 turns, and the cross-sectional area of 3.243x10⁻⁴ m².

Table 4. Objectives of the Wireless FM Microphone Experiment

Design and measure a 1.0 μH solenoidal air-core inductor

Analyze and build an audio microphone amplifier circuit.

Learn about the two conditions for oscillation in a feedback oscillator circuit.

Learn how to analyze a typical RF "LC" oscillator circuit.

Build/debug RF oscillator, then add audio modulation circuit to make a "wireless microphone".

Experiment with radio wave propagation and different polarizations of radiated EM waves.

Table 5. Instruments Used in the Wireless FM Microphone Experiment

Agilent triple DC power supply
Agilent Spectrum Analyzer
Agilent 0 – 100 MHz Digital Oscilloscope
Agilent 0 – 20 MHz Function Generator
Portable FM radio (Walkman style or boom box style)
Tektronix Curve Tracer
Agilent triple DC power supply

The resulting coil inductor was measured in series with a 10 nF capacitor as in Figure 8 to obtain the resonant frequency of 1.52 MHz, which shows the inductance of the coil to be 1.096 μ H.

3.3 The Audio Amplifier Construction and Measurement

The audio amplifier circuit is shown in Figure 9. Please note that this amplifier uses the simplest possible biasing circuit for simplicity, though a 3-resistor bias network would give better Q-point stability with respect to beta variation. The beta value of the transistor is measured using a curve tracer to be 160. The DC bias Q point is then calculated as follows:



Figure 8. Measuring the Inductance Using Series Resonance

$$I_{bq} = \frac{9V - 0.7V}{470k\Omega} = 17.66 \,\mu \,A$$

$$I_{cq} = \beta I_{bq} = 2.826 \,\text{mA}$$

Therefore, $V_{ceq} = 9V - I_{cq} \times 560\Omega = 7.418 \,V$

The measurement of our circuit gives the Q point of Vceq = 7.46V and Icq = 2.75mA, which are close to the theoretical values of 7.42V and 2.83mA. The gain of the amplifier can be calculated as follows.

$$r_{pi} = \frac{26mV}{I_{bq}} = 1.472k\Omega$$
$$Av = \frac{V_{O}(t)}{V_{I}(t)} = \frac{-\beta R_{C}}{r_{pi}} = \frac{-160 \times 560\Omega}{1.472k\Omega} = -60.87$$

The measured AC voltage gain is measured by replacing the microphone with a function generator at 1 kHz and 10 mV amplitude is -52, which is reasonably close to the calculated value of -60.9, considering the very simple small-signal BJT model that was used.

3.4 The Analysis of the Radio Frequency Oscillator Circuit

The RF oscillator circuit is shown in Figure 10. The beta value of transistor Q2 was measured on a curve tracer to be 160. The value of the DC quiescent operating point or Q-point of I_{cq2} and V_{ceq2} are calculated as follows.



Figure 9. The Audio Amplifier Stage and its AC Model.

$$I_{b2q} = \frac{9.0V - 0.7V}{10k\Omega + (160 + 1) \times 1k\Omega} = 48.54 \mu A$$
$$I_{c2q} = \beta I_{b2q} = 160 \times 48.54 \mu A = 7.766 mA$$
$$V_{ce2q} = 9.0V - (160 + 1 \times 1k\Omega \times I_{b2q} = 1.185V)$$

The measured value of the Q point was $I_{c2q}=7.73$ mA and $V_{ce2q}=1.27$ V, which agreed closely with the theoretical values.

The AC model of the RF oscillator circuit was constructed as shown in Fig. 10 assuming that the two 0.001 μ F capacitors, Cbypass1 and Cbypass2 act as short circuits at 34 MHz, since the impedance magnitude of a 0.001 μ F capacitor at 34 MHz is less than 5 ohms. Note that Cbypass2 should not be made any larger than 0.001 μ F, since we do not want it to short out the audio signal from the microphone, and prevent it from reaching the base of the RF oscillator transistor, where it is used to slightly vary the reverse bias voltage across the BC junction of the oscillator stage back and forth at an audio frequency rate.

The base of BJT Q2 is grounded in the radio frequency ac model of our RF oscillator circuit, and so is the top side of the parallel resonant "tank" circuit formed by Lx and Cx, which had been connected to Vcc. Thus the BJT functions as a "common base" amplifier. The input signal voltage is delivered to the emitter terminal (E), creates a base current $i_{b2}(t) = -v_E(t)/r_{\pi 2}$, and the amplified output appears at the collector terminal (C). Note that the output is fed back to the input via a frequency-selective feedback network that consists of Re2, Cfdbk, Lx, and Cx.

This feedback network completes a "feedback loop" that allows oscillations to build up only at a particular frequency "f". This frequency of oscillation is primarily set by the parallel resonant tank circuit formed by Lx and Cx. Because admittances add in parallel, recall that this parallel resonant circuit exhibits the impedance

$$Z_{RxCx} = \frac{1}{j2\pi fCx + \frac{1}{j2\pi fLx}}$$

This impedance becomes infinite when its denominator is set to zero. This occurs at the frequency where the positive imaginary capacitive admittance which is increasing with frequency cancels the negative imaginary inductive admittance which is decreasing with frequency, i.e., $j2\pi fCx + \frac{1}{j2\pi fLx} = 0$.



Figure 10. The 34 MHz RF Oscillator Circuit and its AC Model.

The frequency at which infinite impedance occurs is called the resonant frequency, which in this case is

fres =
$$\frac{1}{2\pi\sqrt{LxCx}} = \frac{1}{2\pi\sqrt{10^{-6} \cdot 22 \times 10^{-12}}} = 33.93 MHz$$



Figure 11. Amplifier/Feedback Network and Loop Gain Measurement.

At frequencies below the resonant frequency "fres", the right term in the denominator of the Z_{RxCx} dominates, and the circuit exhibits a relatively low (inductive) impedance. Likewise at frequencies above the resonant frequency, the left term in the denominator of the Z_{RxCx} expression dominates, and the circuit exhibits a relatively low (capacitive) impedance. Thus this circuit serves to divert to ground signals of frequencies other than "fres". Therefore, only noise at the frequencies near fres are passed around the feedback loop, allowing a significant oscillation to build up.

A general sinusoidal feedback oscillator consists of an amplifier and some sort of an external feedback network, as shown in Figure 11. This general sinusoidal oscillator circuit will oscillate at a frequency "f" if and only if a frequency "f" can be found that satisfies the following two conditions for oscillation, known as the Barkhausen conditions:

The magnitude of the voltage gain around the loop (loop gain) is greater than 1, so that noise at frequency "f" that is present due to the power-up transient will be amplified to higher and higher levels as it circulates around the frequency-selective feedback loop $|A(f)\beta(f)| > 1.0$.

The phase shift around the loop must be an integral multiple of 2π radians, so that the fed back sinusoidal signal will add "in phase" (constructively) with the signal already at the input,



Add Rx in series with Lx to model the real inductor more accurately.

Figure 12. Breaking the Feedback Loop in the AC Model of Figure 10.

and then oscillations can then build up. Thus we require that the phase angle $[A(f)\beta(f)] = n2\pi$, where n is any integer. The loop gain can be measured by breaking the feedback loop and inserting a sinusoidal test source on the input side of the break, and then measuring the voltage gain around the loop, as shown in Figure 12.

Returning to the AC model of our RF oscillator circuit in Figure 10, the feedback loop can be probed most conveniently by disconnecting Cfdbk from the collector of Q2 and inserting a sinusoidal test source, Vin(t). This is shown in Figure 10. To model the inductor more closely, we should include a small series wire resistance, Rx, in series with Lx. We shall take this resistance to be 0.25 Ω when analyzing this circuit. This prevents the impedance of the series resonant circuit from actually going to infinity at its resonant frequency, though it will still go to a very high value.

We may now write the node voltage equations at the emitter node (Ve) and the collector node (Vout) in terms of the Laplace complex frequency variable "s"

$$\frac{\text{Vout}}{\text{sL}_{x} + \text{R}_{x}} + \text{Vout} \cdot \text{s} \cdot \text{Cx} - \frac{\beta_{2} \cdot \text{Ve}}{r_{\pi 2}} = 0$$
$$\frac{\text{Ve}}{r_{\pi 2}} + \frac{\text{Ve}}{\text{R}_{e2}} + (\text{Ve} - \text{Vin}) \cdot \text{s} \cdot \text{C}_{\text{fdbk}} + \frac{\beta_{2} \cdot \text{Ve}}{r_{\pi 2}} = 0$$

Then eliminating Ve, we can show that the loop gain is given by

Loop Gain =
$$\frac{\text{Vout}(s)}{\text{Vin}(s)} = \frac{sC_{\text{fdbk}}R_{e2}\beta_2(sL_x + R_x)}{(s^2C_xL_x + sC_xR_x + 1)(sC_{\text{fdbk}}R_{e2}r_{\pi 2} + r_{\pi 2} + (\beta_2 + 1)R_{e2})}$$

Replacing s by $j2\pi f$ and plotting the magnitude and angle of the loop gain, and substituting in the component values (with the series resistance of the coil modeled by $Rx = 0.25 \Omega$) we can show that the loop gain phase shift passes through 0 degrees at f = 34.35146 MHz, where the loop gain amplitude is found to be 40.3, which is considerably greater than unity, so the circuit will oscillate dependably.



Figure 13. 34 MHz Microphone and its Output Voltage Waveform.

3.5 The Construction and Testing of the FM Wireless Microphones

Figure 13 shows the breadboard construction of the 34 MHz wireless phone and its fundamental frequency waveform which is close to 34 MHz. The waveform is quite distorted to produce large harmonics. The coil may be pulled apart or pushed back together to alter the resonant frequency as desired.

The radiation spectrum of the wireless phone was measured with a spectrum analyzer and a 12" wire antenna connected to the spectrum analyzer input and placed near the wireless FM microphone as shown in Figure 14.

Since the transmitting antenna of the wireless phone is also a 12" wire, the received 2^{nd} harmonic of 67.5 MHz was 10 dB above the fundamental frequency of 33.7 MHz. The 3^{rd} harmonic of 100.5 MHz was 3 dB below that of the 2^{nd} harmonic. The 4th harmonic was about equal in strength to the 3^{rd} harmonic while the higher harmonics rapidly fell off in amplitude.



Figure 14. Spectrum of the Received Signal from the 34 MHz Wireless Microphone.

This FM receiver is able to pick up speech clearly from 25 or more feet away from the microphone. The frequency of oscillation was not very stable since it is determined by Lx and Cx. Moving a hand near the coil would detune the oscillation by several hundred kiloHertz. The maximum range of wireless microphone circuit was only about 50 feet.

Different antenna orientations and antenna lengths were explored with this setup. For example, a quarter-wave monopole of about $\frac{3}{4}$ meter in length was found to be a good antenna length for the 3-meter wavelength, 102 MHz 3^{rd} harmonic signal.

Once the circuit was found to work successfully at 34 MHz, the students were asked to try to make their wireless microphone circuit oscillate directly at 102 MHz. This flys in the face of conventional wisdom, as most professors will tell their students that prototyping breadboards will "not work at RF"! By replacing the 1 nH inductor (Lx) with a 1.5" loop of wire, most students found that it was possible to make their circuit oscillate directly at 102 MHz, despite our concerns about using an inexpensive 2N3904 switching transistor as the oscillator and a standard prototyping breadboard with its relatively high trace inductance and intertrace capacitance. The modified circuit and its output spectrum is shown in Figure 15. The frequency setting inductor "Lx" is the short, vertical blue loop of at the right of the breadboard. The loop length could be adjusted to tune the oscillating frequency to the FM band from 88 MHz to 108 MHz. The received electromagnetic wave spectrum in Figure 15 shows that the fundamental frequency of the wireless phone is indeed now 102 MHz. Perhaps the most interesting lesson that students take away from this lab experiment is "what a difference a short loop of wire makes!" Since such a short loop of wire has become such an important frequency-setting inductor in this circuit, it is clear that all other interconnecting wires (where inductors are not wanted, must be kept considerably shorter than this 1.5" loop!



Figure 15. A 102 MHz Wireless Microphone Construction and Its Radiation Spectrum.



Figure 16. The 3-Inverter Ring Oscillator Circuit and Its Breadboard Consstruction.

4 Experiment #3: Selection of Bypass Capacitors for a DC Power Supply

Selection of bypass or decoupling capacitors has been discussed extensively in the electronic design community⁸⁻¹². Although the topic has received great attention in graduate schools, it may not be addressed adequately in undergraduate education. This knowledge needs to be transferred to undergraduate computer engineering students who would then apply the knowledge in their digital system designs after they start their engineering careers.

This student laboratory experiment investigates how different values of bypass capacitors on a DC power bus can reduce DC power bus noise voltage fluctuations and radiated emission from the bus. The experimental platform is a 3-inverter ring oscillator built using a 74HC00 digital IC on a breadboard, as shown in Figure 16. The oscillator is powered by a 5V DC power supply with bypass capacitors. Noise voltage fluctuations on the DC power bus is observed on an oscilloscope, and radiated emissions up to 600 MHz are observed on a spectrum analyzer, as different bypass capacitor combinations are placed across the power and ground pins.

4.1 Objectives and the Equipment List

The main objectives of this experiment, as listed in Table 6, are to learn about the operation of ring oscillators, study the effects of using different bypass capacitors on DC power bus noise and also on conducted emissions.

Table 7 lists the instruments and components needed to perform this lab. The only uncommon instrument is the spectrum analyzer.

Learn the ring oscillator, and propagation time, rise time, and fall time measurements.
Investigate effects of various types of bypass capacitors on radiated emissions.
Investigate effects of various types of bypass capacitors on Vcc(t) dc power bus waveform.
Learn how to use the spectrum analyzer/tracking generator to model a real capacitor at high
frequencies

Table 0. Objectives of the Dypass Capacitor Experime	Table 6.	Objectives	of the Bypa	ss Capacitor	Experiment
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Table 7. Instruments and Components for the Bypass Capacitor Experiment

Agilent 54624A 100 MHz Digital Oscilloscope
Agilent E4402B Spectrum Analyzer (100 Hz – 3 GHz)
Agilent E3631AVariable Triple Output DC Power Supply
Prototyping Breadboard
Assorted capacitors with short leads (10 µF, 0.1 µF, 0.001 µF (1 nF), 100 pF)
74HC04 High Speed CMOS Hex Inverter

4.2 Construction and Measurement of the 3-inverter Ring Oscillator

The output of an N-inverter ring oscillator, where N is odd and greater than unity, will toggle after N gate delays (N*t_{prop}) and therefore produce a rectangular waveform. It takes N*t_{prop} seconds for a change of state at the output to propagate back to the output, causing the output to change state. The period of the oscillation is therefore $T_{osc} = 2N*t_{prop}$ and the oscillation frequency is $f_{osc} = 1/T_{osc} = 1/(2N*t_{prop})$. Figure 16 shows a 3-inverter ring oscillator. The rightmost inverter is not part of ring and is used to buffer the output so that the capacitive loading on the last gate in the ring would not slow down the final inverter stage in the ring.

The measured frequency of the oscillator in Figure 16 was 35.34 MHz. The average gate delay is therefore 4.72 ns=1/(6*35.34MHz). The students are next requested to verify that an N = 4 ring will not oscillate, and that an N = 5 ring will oscillate near the lower frequency of 1/(10* t_{prop}).= 1/(10*4.72 ns) = 21.19 MHz.

4.3 Effect of DC Power Bus Bypass Capacitors on Radiated Emissions

The radiation spectrum of the 3-inverter ring oscillator was measured by 12" monopole wire antenna that was placed about 2 feet away from the circuit. Three different bypass capacitors were used to evaluate their effectiveness in reducing radiation and their measured radiation spectrum up to 600 Mhz are presented in Figure 17.

As can be seen from Figure 17, there is about 10 dB reduction in radiation between the use of the 100 pF capacitor and 1 nF capacitor from -23 dBm to -32.5 dBm. However the difference between radiation reduction of using the 1 nF capacitor and 4.7 μ F one was less than 1 dB. This suggests that capacitors with larger values are less effective than those of smaller values because of larger inductive effects and lower resonant frequencies of the larger capacitors.

4.4 Effect of DC Power Bus Bypass Capacitors on Power and Ground Noises

Figure 18 shows power and ground bus noises with three different decoupling capacitors. Higher valued capacitors permit more power and ground noise glitches, which significantly slow the rise and fall times of the output signal from the ring oscillator. This explains why lower radiated emissions are observed from circuits with higher bypass capacitor values.



Figure 17. Radiation Spectrum of the Oscillator with Various Bypass Capacitors.



Figure 18. Observed Power Bus Noise Voltage, Vcc(t), with Different Bypass Capacitors.

The students are surprised to find that the relatively small 100 pF capacitor does a better job reducing Vcc(t) fluctuations than the larger bypass capacitors.

To investigate why this is so, the students are next requested to use the spectrum analyzer with its associated tracking generator to determine the equivalent lead inductance as well as the self-resonant frequencies of various bypass capacitors with various lead lengths. Typical results are shown in Table 8.

The results contained in this table permits the student to explain why the 100 pF bypass capacitor is more effective than the larger bypass capacitors in this circuit. Because this particular circuit is a simple 35-MHz ring oscillator, the power bus noise occurs at harmonically related noise frequencies of 35 MHz, 70 MHz, 105 MHz. Thus the power bus noise is quite high frequency in this lab experiment. Since a bypass capacitor stops acting like a capacitor above its self resonant frequency, and instead starts resembling an inductor, Only the 100 pF capacitor (fres = 80 MHz) is effective in bypassing the first and second harmonic noise components. The 1 nF and larger bypass capacitors have a self-resonant frequency that is below even the first harmonic noise components, so they cannot do a very good job bypassing the noise on the power bus.

Marked Value	Lead Length (cm)	Arelative (dB)	fx (MHz)	fres (MHz)	C1	L1
100 pF	0.5	1.54	40.5	80	102 pF	10.6 nH
1 nF	0.5	1.73	4.58	27.9	0.97 nF	33.5 nH
1 nF	2	1.79	4.58	22.9	0.99 nF	48.7 nH
0.1 uF	0.5	5.6	0.103	4.27	0.100 uF	13.9 nH
0.1 uF	2	7.26	124	3.55	0.107 uF	18.8 nH
0.33 uF	0.5	17.8	0.200	1.45	0.245 uF	49.1 nH
0.33 uF	2	12.1	0.125	1.15	0.200 uF	96.4 nH

Table 8. Typical Bypass Capacitor Values With Different Lead Lengths and Their Measured Capacitance (C1), Equivalent Lead Inductance (L1), and Measured Self-Resonant Frequency (fres)

Certainly the most important message that the student takes away from this lab is "Bigger is not always better!" It should be pointed out, however, that larger bypass capacitors in parallel with smaller ones would be needed in a digital system with irregular switching rates, which gives rise to Vcc(t) noise at many different frequencies, both low and high.

5 Conclusions

Three RF and EMC experiments are described and analyzed in this paper. Measurement results from them are included and discussed. The experiments are inexpensive to implement and flexible enough for instructors to create different versions or complexities to adapt to the need of a class on EMC, SI or RF.

These three experiments work together to expose the students to many common topics in EMC such as common-mode currents and their suppression with a common-mode choke, switching DC power supply design, conducted emissions, antenna length and polarization, harmonics of a rectangular waveform, mutual and self inductances, audio amplifiers, radio frequency oscillators, FM modulation, effectiveness of bypass capacitors, and the self-resonant frequency of a capacitor.

The actual laboratory handouts presented to the students for these three labs, as well as video presentations of these experiments by one of the authors of this paper are available at the following website: www.rose-hulman.edu/~hoover.

6 Bibliography

- 1. C.R. Paul, "Establishment of a university course in electromagnetic compatibility (EMC)," IEEE Transaction on Education, Vol. 33, pp.111-118, Feb. 1990.
- 2. Robert M. Nelson, "EMC education at North Dakota State University," 1992 IEEE International Symposium on EMC, pp.164-167, 1992.
- 3. Donald D. Weiner, "EMC education at Syracuse University," 1992 IEEE International Symposium on EMC, pp.168-170, 1992.
- 4. C.W. Trueman, "An electromagnetics course with EMC applications for computer engineering students," IEEE Transaction on Education, Vol. 33, pp. 119-128, Feb. 1990.
- 5. Y. Zhao and K. Y. See, "A practical approach to EMC education at the undergraduate level," IEEE Transaction on Education, Vol. 47, No. 4, pp. 425-429, November 2004.
- 6. Intel's Signal Integrity for High Speed Circuits, http://www.intel.com/education/highered/signal/index.htm.
- 7. IEEE EMC Education Manual, IEEE EMC Society available from its website.
- 8. Tamara Schmitz and Mike Wong, "Choosing and Using Bypass Capacitors," Application Note AN1325.0, Intersil, 2007.
- 9. Istvan Novak, "Comparison of Power Distribution Network Design Methods: Bypass Capacitor Selection Based on Time Domain and Frequency Domain Performances," DesignCon 2006.
- 10. Lattice Semiconductor, "Power Decoupling and Bypass Filtering for Programmable Devices," Technical Note TN1068, May 2004.
- 11. Mark Alexander, "Power Distribution System (PDS) Design: Using Bypass/Decoupling Capacitors," Application Note XAPP623, Xilinx, 2005.
- James L. Knighten, Bruce Archambeault, Jun Fan, Giuseppe Selli, Liang Xue, Samuel Connor, and James L. Drewniak, "PDN Design Strategies: II. Ceramic SMT Decoupling Capacitors – Does Location Matter?," Newsletter of the EMC Society of the IEEE, Winter 2006.