

# REDUCTION OF POWER SUPPLY EMI EMISSION BY SWITCHING FREQUENCY MODULATION

by

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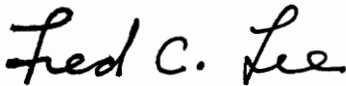
in

Electrical Engineering

APPROVED:



Dr. D. Y. Chen, Chairman



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Committee Chairman: D.Y.Chen  
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## **(ABSTRACT)**

The effect of PWM frequency modulation on power supply conducted EMI noise emission is investigated. Significant reduction of emission is possible with PWM frequency modulation scheme.

A forward converter is used to verify the effectiveness of the scheme. A guidance of parameter selection for noise reduction is given.

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## Table of Contents

<b>Chapter I INTRODUCTION .....</b>	<b>1</b>
<b>Chapter II MEASUREMENT OF EMI AND GOVERNMENT REGULATIONS .....</b>	<b>3</b>
2.1 EMI Measurement Setup .....	3
2.1.1 LISN .....	3
2.2 Government EMI Regulations .....	5
2.3 Spectrum Analyzer .....	7
2.3.1 Resolution Bandwidth .....	7
2.3.2 Video Bandwidth .....	7
2.3.3 Sweeping Time .....	13
<b>Chapter III EMI NOISE IN POWER SUPPLY .....</b>	<b>14</b>
3.1 Differential-Mode And Common-Mode Noise .....	14
3.2 Conducted D.M. Noise In A Switching Power Supply .....	14
3.3 Conducted C.M. Noise In A Switching Power Supply .....	17
3.4 Radiated Noise .....	18
3.4.1 Differential Mode .....	18
3.4.2 Common Mode .....	19
<b>Chapter IV EFFECT OF FREQUENCY MODULATION ON NOISE SPECTRUM .....</b>	<b>22</b>
4.1 Spectrum Of Single-Tone Modulation .....	22
4.2 $\beta$ Value .....	25
4.3 Carson's Rule .....	25
4.4 Spectrum Of Pulse Train .....	31
<b>Chapter V VERIFICATION OF POWER SUPPLY NOISE .....</b>	<b>38</b>
5.1 Description Of Testing Circuit .....	38
5.1.1 PWM-Frequency Modulation Circuit .....	40

5.2 Measurement Results ..... 44

5.3 Parameters Affecting Performance ..... 51

5.3.1  $\beta$  Value ..... 51

5.3.2  $\Delta f$  Value ..... 59

5.3.3  $f_m$  Value ..... 59

5.3.4 Table Summarizing The Effect Of Parameter Change ..... 60

5.4 Possible Adverse Effects Caused By FM Of PWM Frequency ..... 60

**Chapter VI CONCLUSIONS AND FUTURE RESEARCH ..... 63**

6.1 Conclusions ..... 63

6.2 Future Research ..... 64

**REFERENCES ..... 66**

**VITA ..... 68**

# **Chapter I**

## **INTRODUCTION**

Electro-Magnetic Interference ( EMI ) emission is always of grave concern for power electronic circuit designers. Due to rapid switching of high current and high voltage, interference emission is a serious problem in switching power circuits. The emission could couple through the power line or radiate through the air to the victim receiver. Many products fail to make it to the market because of failure to comply with the government EMI regulations. Numerous companies have cited EMI problems as the culprit for the delay in their product introduction. This problem should become even more serious as government EMI regulations tighten even further in the future.

EMI noise reduction is generally accomplished by three means: suppression of noise, isolation of noise coupling path, and filter/shielding. In this thesis, another means of EMI reduction is proposed and investigated. By modulating the PWM frequency of a switching power supply, the noise emission spectrum is suppressed. The total power of the emission remains the same, but the amplitude is suppressed. This makes it easier to pass EMI regulations.

In the thesis, measurement of EMI noise and government regulations are briefed in Chapter II. This serves as the common ground for assessing the effectiveness of the proposed scheme. Chapter III gives a discussion on the noise sources and coupling paths in a typical switching power supply circuit. Chapter IV describes the spectrum of a frequency-modulated waveform. Mathematical effect of frequency-modulation is discussed in detail. The results serve as the theoretical foundation for the experiments to be discussed in Chapter V. In Chapter V, detailed EMI measurements resulting from a frequency-modulated power supply circuit will be investigated. Practical impact of the

proposed means for EMI reduction will be assessed. A guidance of parameter selection to minimize the noise is given. Chapter VI concludes the thesis and outline possible future work in the area.

## **Chapter II**

### **MEASUREMENT OF EMI AND GOVERNMENT REGULATIONS**

In this chapter, a brief description of the EMI noise measurement and the government EMI specifications will be given. The details of the measurement setup is quite involved and can be found from federal agencies documents, and therefore will not be repeated here. Only those which are pertinent to this thesis work are outlined in this chapter.

#### **2.1 EMI Measurement Setup**

Figure 2.1 shows the circuit diagram for EMI measurement. The dotted box shows a line impedance stabilizing network ( LISN ). The noise voltage measured across the  $50\ \Omega$  resistor, usually displayed in frequency domain, is by definition the conducted EMI emission of the switching power circuit. In this chapter, the compliance measurement and its setup are the main concern. FCC ( Federal Communications Commission ) and VDE ( Germany standard ) use the measurement setup requirement of the International Special Committee on Radio Interference ( CISPR ).

##### **2.1.1 LISN**

In the measurement setup, a line-impedance stabilization network ( LISN ) is used for conducted EMI emission. The purpose of a LISN is to prevent the line impedance  $Z_s$  from affecting the measurement result, so that the result is repeatable. The values of the reactive components inside LISN are such that for line frequency,  $L$  is essentially

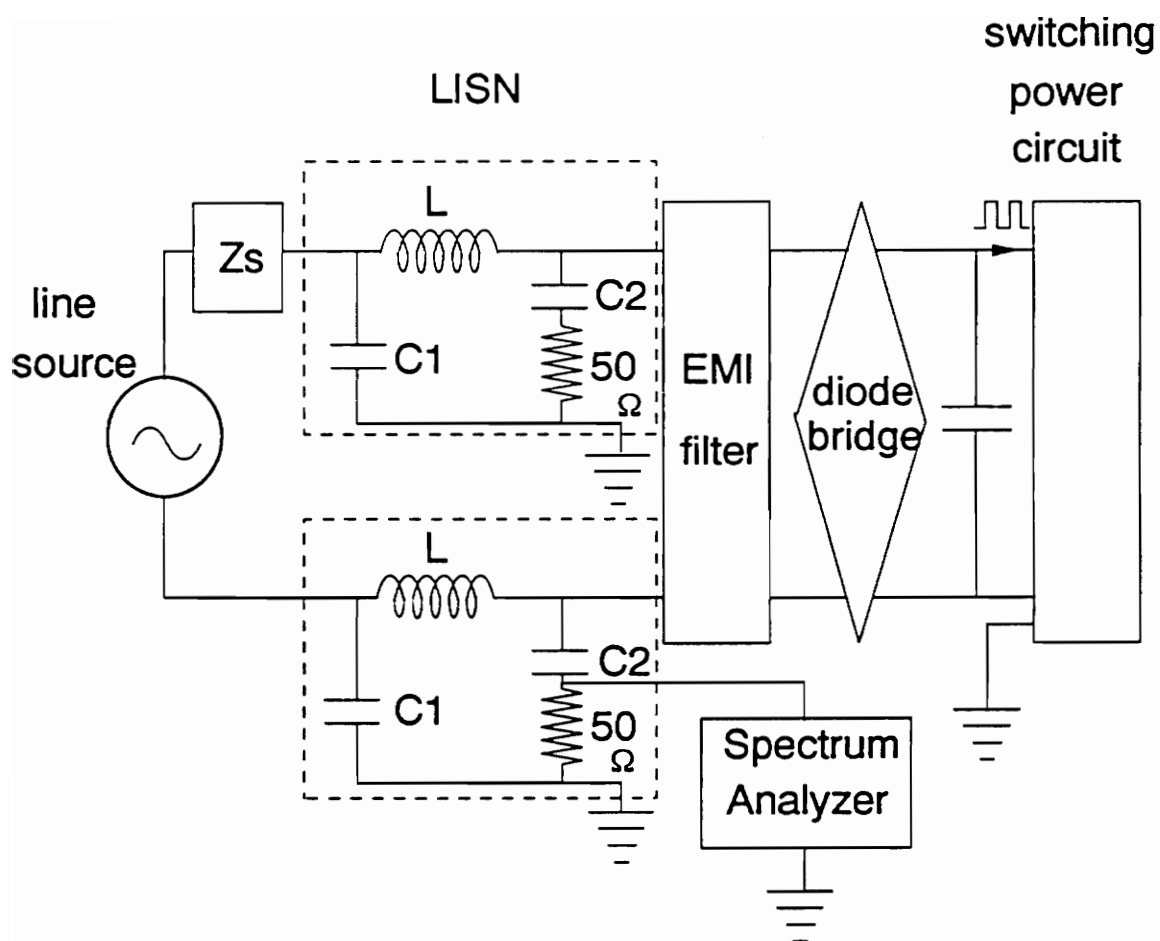


Figure 2.1 Conducted EMI Test Circuit

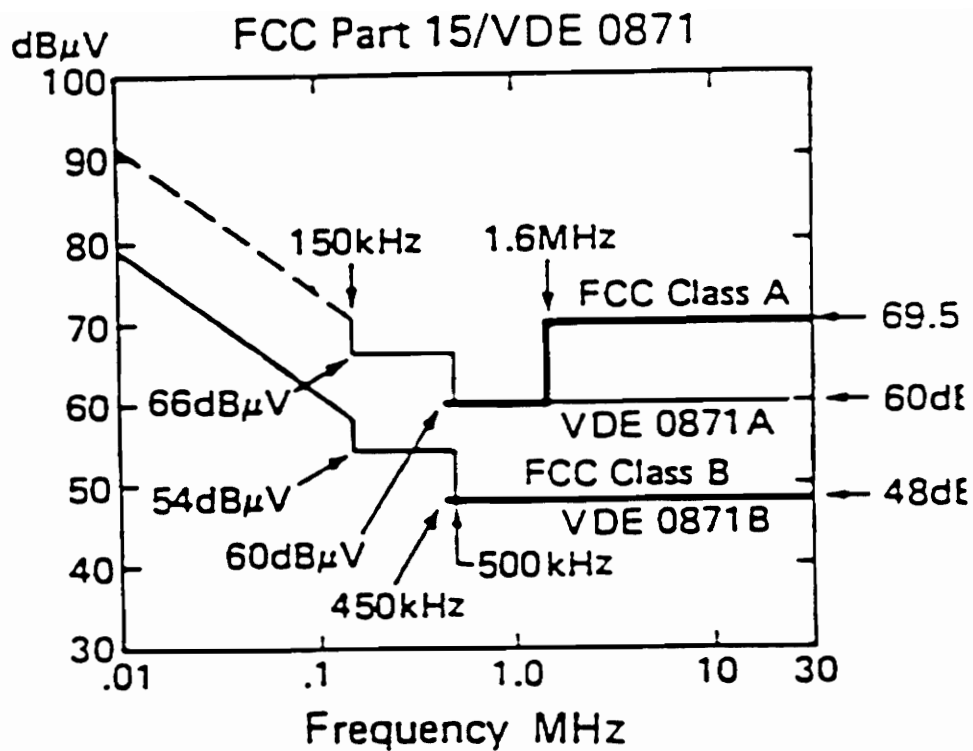
shorted, and C1 and C2 are essentially open. For EMI noise frequency, however, L is essentially open, and C2 is essentially shorted. In other words, line frequency power goes through LISN unaffected, and noise generated by the switching circuit sees a  $50\ \Omega$  impedance from each phase to ground. The  $50\ \Omega$  resistance is normally the input impedance of the measuring spectrum analyzer.

## 2.2 Government EMI Regulations

FCC and VDE regulations both are divided into two classes: Class A and Class B. Class A is used for commercial application, and Class B is for residential application.

The compliance measurement should employ a quasi-peak detector and the related measurement bandwidth required by the International Special Committee On Radio Interference ( CISPR ). However, most spectrum analyzers provide peak detector function, not a quasi-peak detector. Because a peak detector usually catches higher amplitude of EMI than a quasi-peak detector does, if a measurement passes the regulations by using a peak detector, it can usually pass an authorized measurement.

Figure 2.2 shows the conducted EMI specification for FCC and VDE regulations. The different measurement bandwidth is required for different frequency range. VDE requires 200 Hz measurement bandwidth for frequency from 10 kHz to 150 kHz, 9 kHz bandwidth for the range from 150 kHz to 30 MHz. FCC requires 9 kHz measurement bandwidth for the frequency range from 450 kHz to 30 MHz. The unit of the vertical axis is  $\text{dB}\mu\text{V}$ , which is defined as  $20\log V_n/\mu\text{V}$ , where  $V_n$  is the noise voltage. For example, if  $V_n = 10\ \mu\text{V}$ , then it is also expressed as  $20\text{dB}\mu\text{V}$ . The frequency up to 30 MHz is conducted EMI measurement range, and the frequency from 30 MHz to 1 GHz is radiated EMI measurement range. In this thesis, only is conducted EMI concerned.



	Frequency	Measurement Bandwidth
FCC:	450 kHz -- 30 MHz	9 kHz
VDE:	10 kHz -- 150 kHz	200 Hz
	150 kHz -- 30 MHz	9 kHz

Figure 2.2 Conducted EMI Specification For FCC & VDE Regulations

## **2.3 Spectrum Analyzer**

Proper setting of spectrum analyzer is essential in evaluation of the EMI noise. This subsection describes the requirements of these key setting that affect the measurement results.

### **2.3.1 Resolution Bandwidth**

The resolution bandwidth ( RBW ), by definition, is the input filter bandwidth of a spectrum analyzer. If the resolution bandwidth is set small, the spectrum is more discrete. If the resolution bandwidth is set large, it allows more frequency components to be received. As a result, the spectrum becomes " fat ", the lower amplitude of the side-band harmonics could be masked by the envelop of the fundamental component. Figure 2.3.1.1 ( a ) and ( b ) show the difference of using different RBWs for measuring the same noise signal. Figure 2.3.1.2 ( a ), ( b ) and ( c ) show the different measuring results by using different RBWs of a frequency-modulated signal.

In spectrum analyzer, there is no direct measurement-bandwidth setup function, but the resolution bandwidth ( RBW ) is related to the measurement bandwidth, and usually they are close to each other, that is, if the resolution bandwidth is set to 10 kHz, the measurement bandwidth can also be considered 10 kHz, so the measurement-bandwidth can be set by setting the resolution bandwidth to meet the measurement requirement.

### **2.3.2 Video Bandwidth**

In spectrum analyzer setup, the concern about the video bandwidth ( VBW ) in spectrum analyzer is another important factor. VBW is used to average the input signal. In

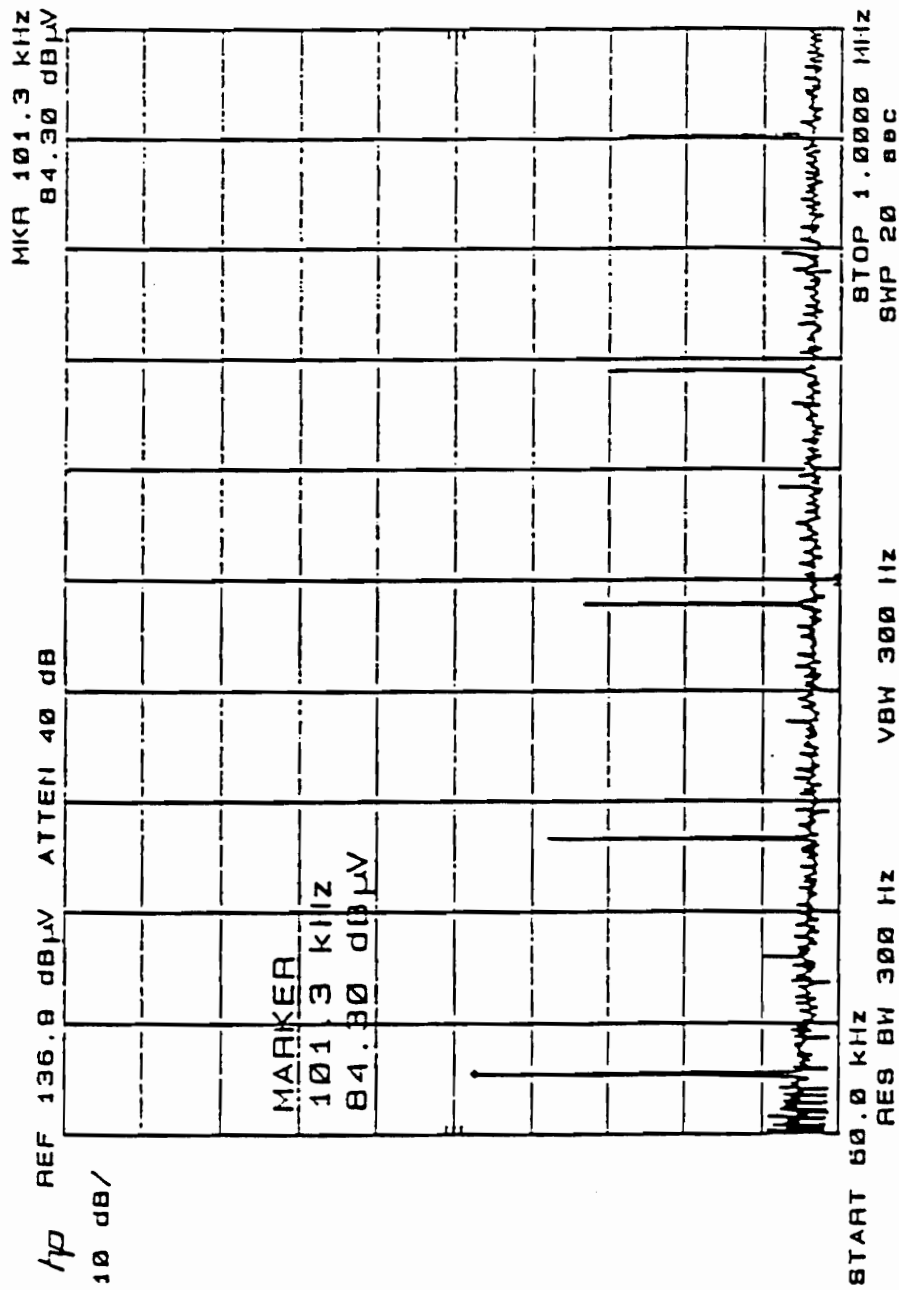


Figure 2.3.1.1 (a) Spectra Measured With 300 Hz RBW

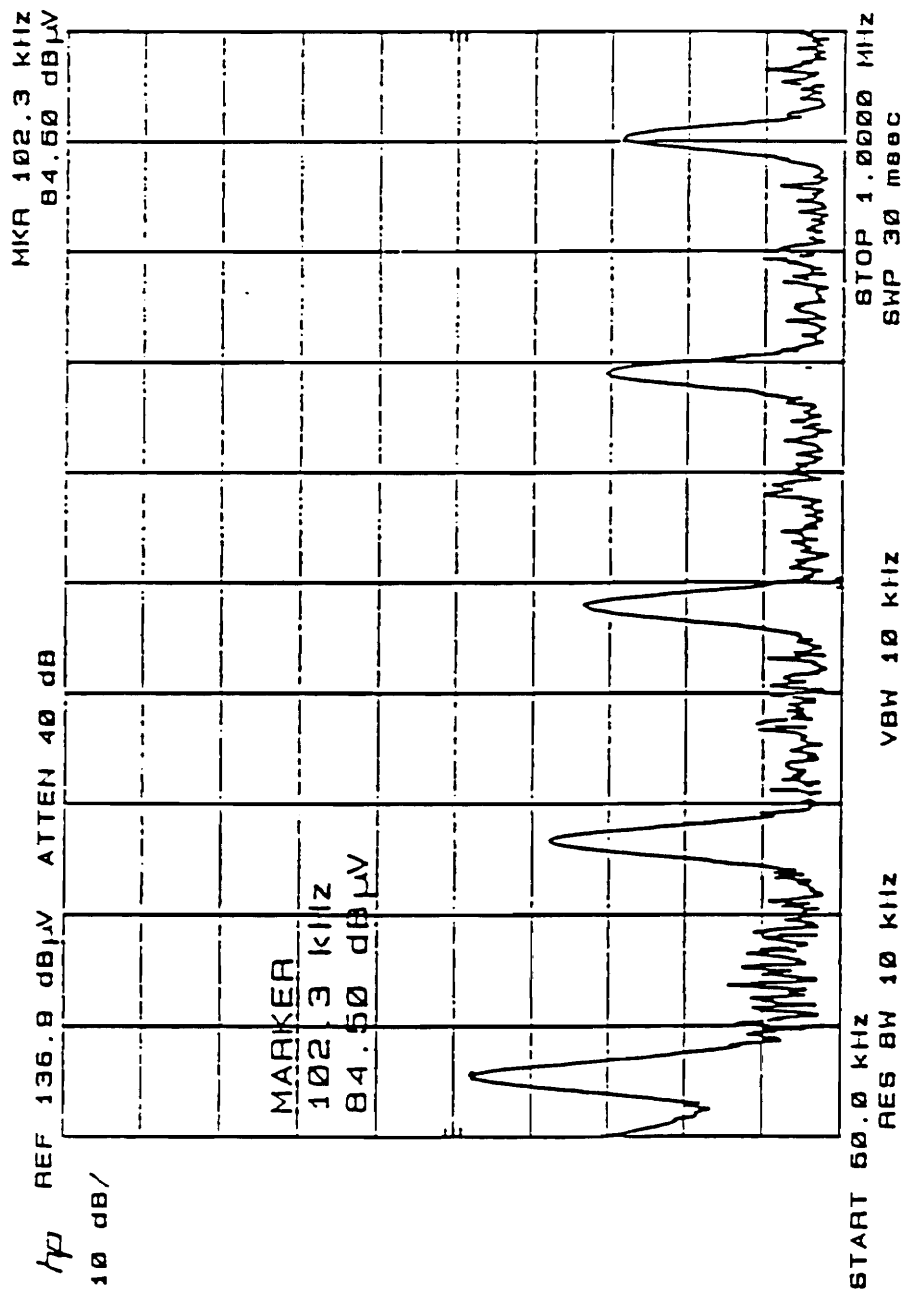
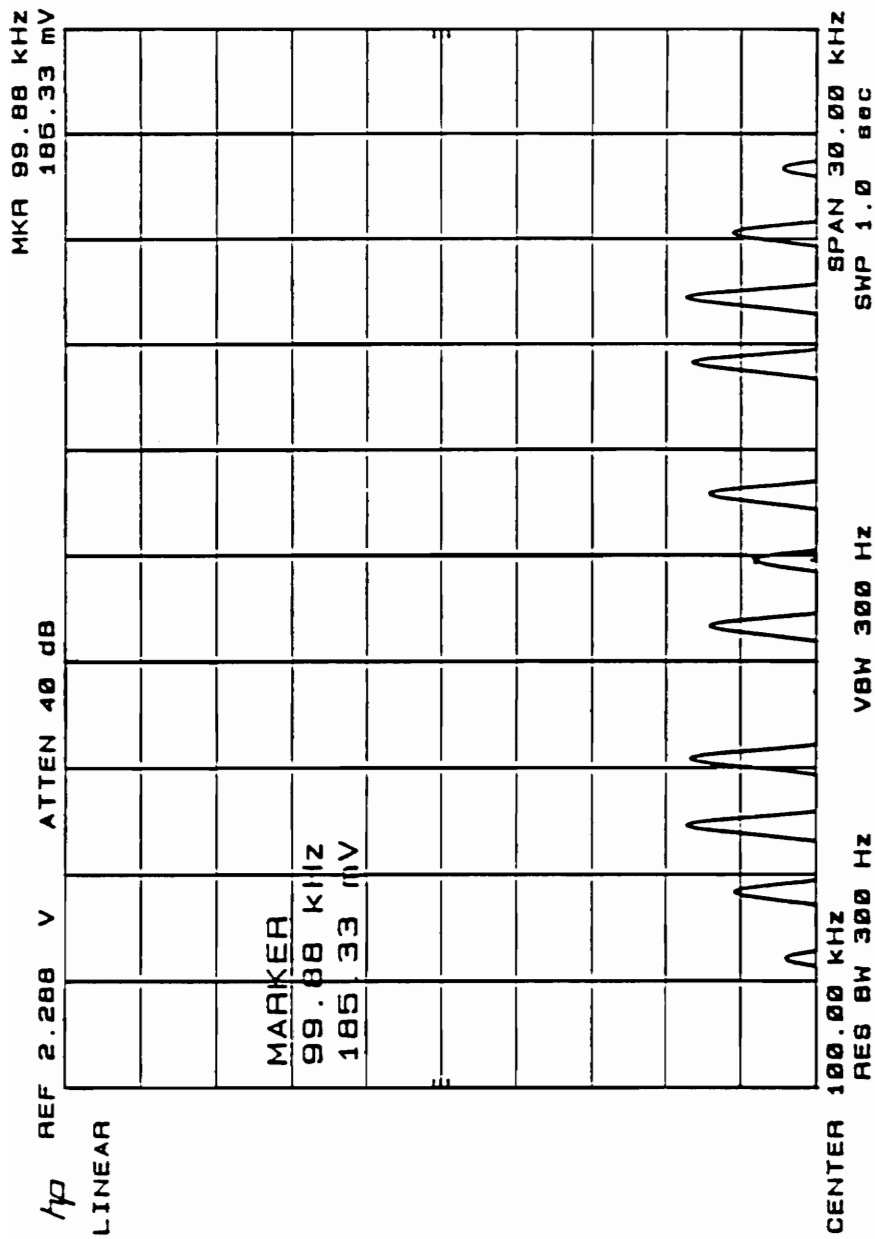
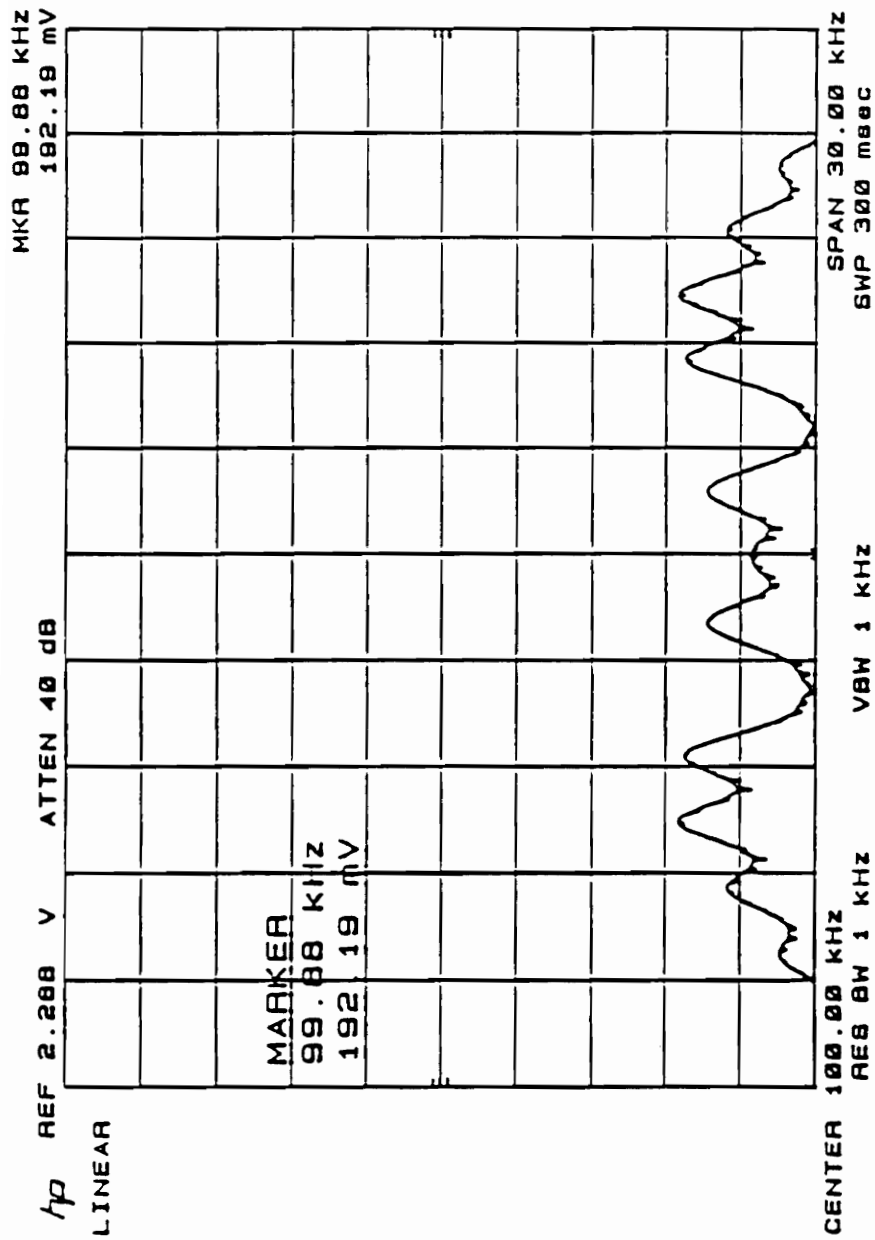


Figure 2.3.1.1 (b) Spectra Measured With 10 kHz RBW



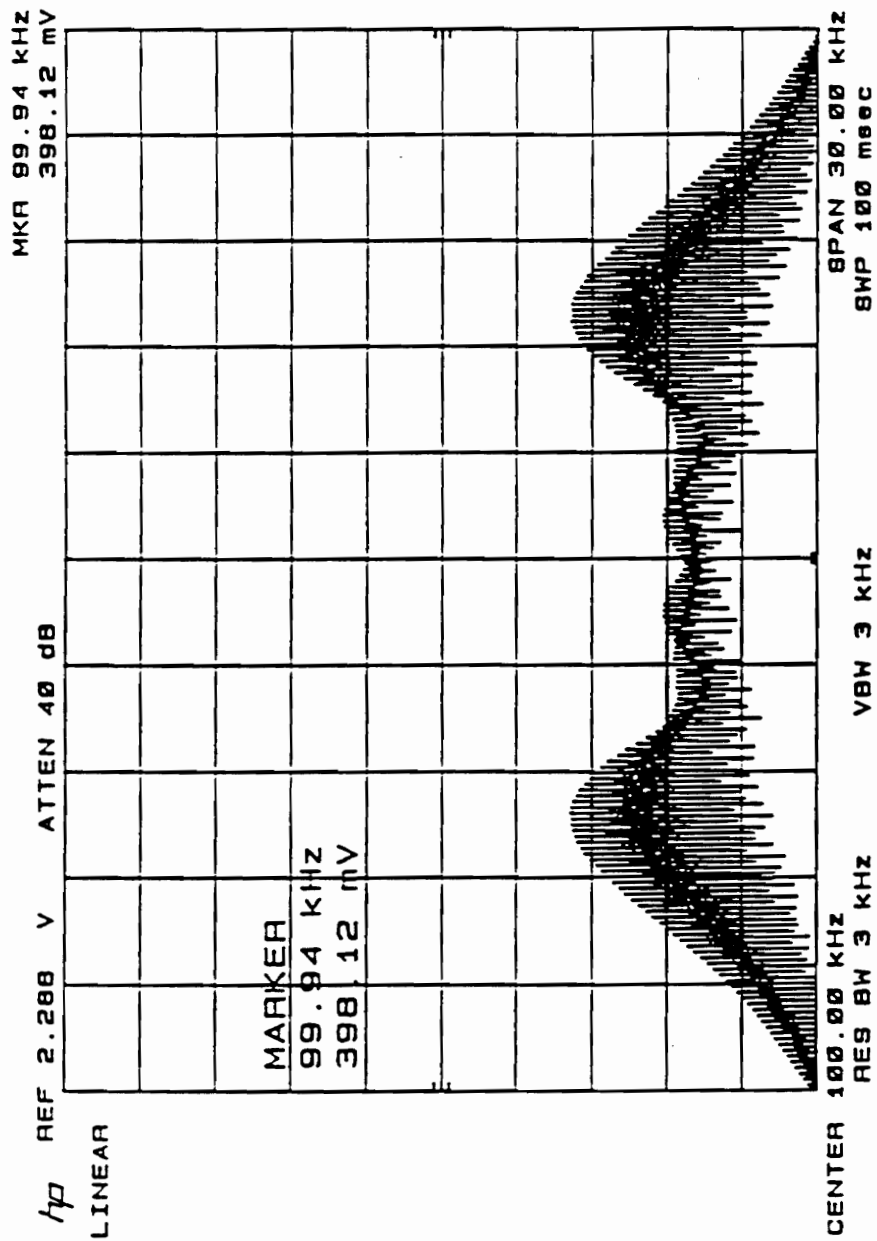
$\Delta f = 10 \text{ kHz}$ ,  $f_m = 2 \text{ kHz}$ , With  $\text{RBW} = 300 \text{ Hz}$

Figure 2.3.1.2 (a) Spectra Of A Frequency Modulated Signal



$\Delta f = 10 \text{ kHz}$ ,  $f_m = 2 \text{ kHz}$ , With  $\text{RBW} = 1 \text{ kHz}$

Figure 2.3.1.2 (b) Spectra Of A Frequency Modulated Signal



$\Delta f = 10 \text{ kHz}$ ,  $f_m = 2 \text{ kHz}$ , With  $\text{RBW} = 3 \text{ kHz}$

Figure 2.3.1.2 (c) Spectra Of A Frequency Modulated Signal

diagnosing measurement, this function can be used to find the fundamental signal which is masked by the noise floor. However, in compliance measurement, the VBW should be set larger or at least equal to the resolution bandwidth, since FCC or VDE requires detecting the noise peak, not the average of the noise. In EMI noise measurement, the VBW is normally set at " automatic mode ", that means it tracks the setting of the RBW.

### **2.3.3 Sweeping Time**

Sweeping time is the time during which the required setting of the frequency range is swept. For the same resolution bandwidth, the longer the sweeping time is, the more frequency components are caught, and the longer working time it takes. If the sweeping time is too short, the amplitude of the measured EMI noise could be lower than that of the actual EMI noise. If it is too long, it unnecessarily increases the working time. In this thesis, the sweeping time is set at " automatic mode ". The sweeping time follows the resolution bandwidth.

## **Chapter III**

### **EMI NOISE IN POWER SUPPLY**

Generally speaking, the EMI noise emission can be divided into two groups: the conducted and the radiated noise emission. The conducted noise emission is coupled through conductors such as wire and parasitic capacitors to the power source. The radiated noise emission is radiated to receivers. In this thesis, the conducted EMI is of primary concern.

#### **3.1 Differential-Mode And Common-Mode Noise**

The conducted EMI emission can be separated into two kinds: The differential mode ( D.M. ) EMI noise and the common mode EMI ( C.M. ) noise. Figure 3.1 shows a three terminal network ( source, load and ground ). From the source or the load side, the two D.M. current are  $180^\circ$  out of phase and do not go through ground terminal. The two C.M. current are in phase and go through the ground terminal. Same thing can be said with regard to voltage. Referring to the ground terminal, when the perturbation occurs, if the voltages on the different points increase or decrease in opposite direction at same time, they are differential mode voltage. If the voltages are in the identical direction, they are the common mode voltage.

#### **3.2 Conducted D.M. Noise In A Switching Power Supply**

Figure 3.2.1 shows a forward converter commonly used in many power supply applications. The two  $50\ \Omega$  resistors are the equivalent noise-frequency impedance of LISN described in Chapter II. D.M. noise is caused by the pulsating transistor current. Part of

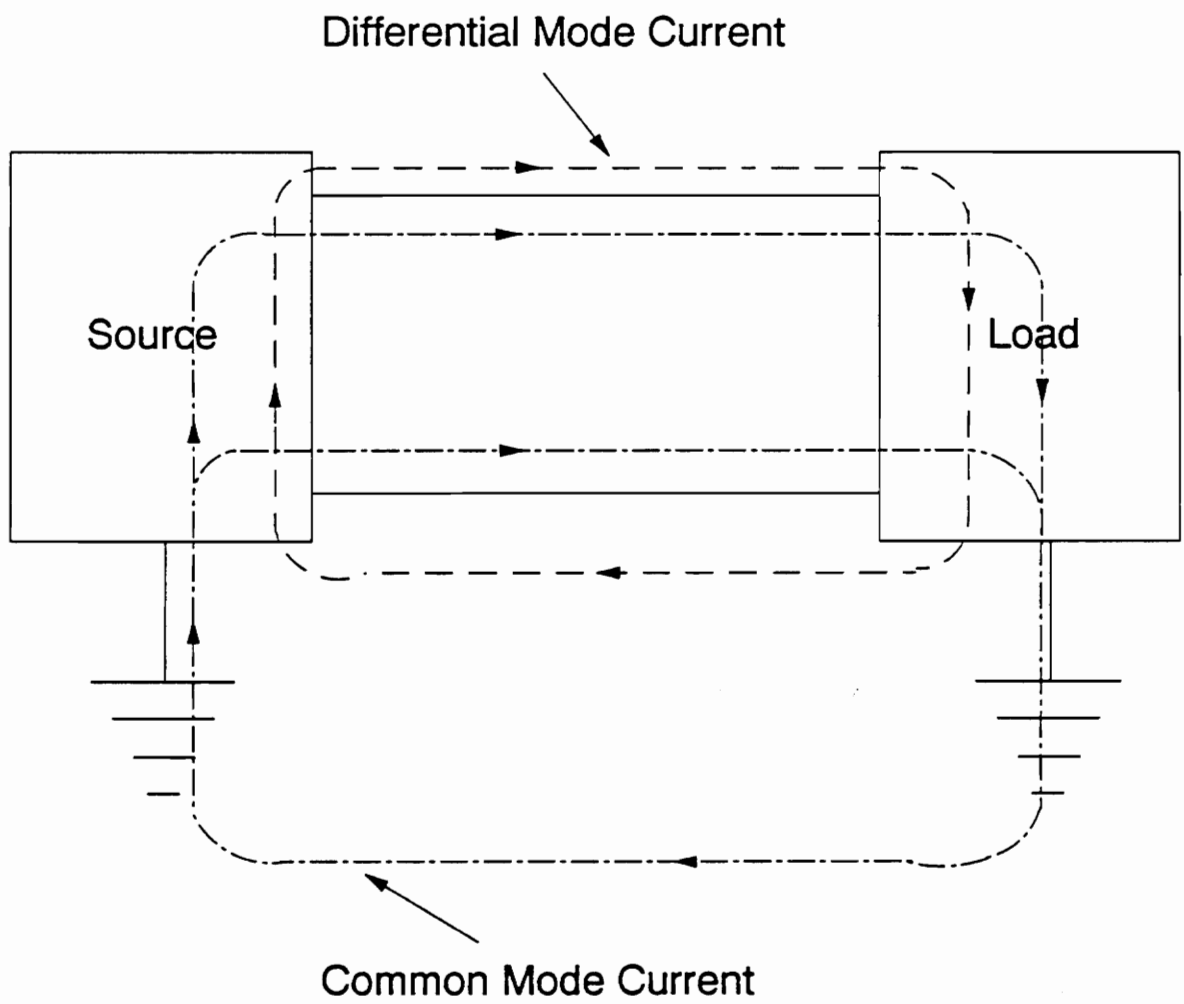


Figure 3.1 D.M. And C.M. Current Depicting

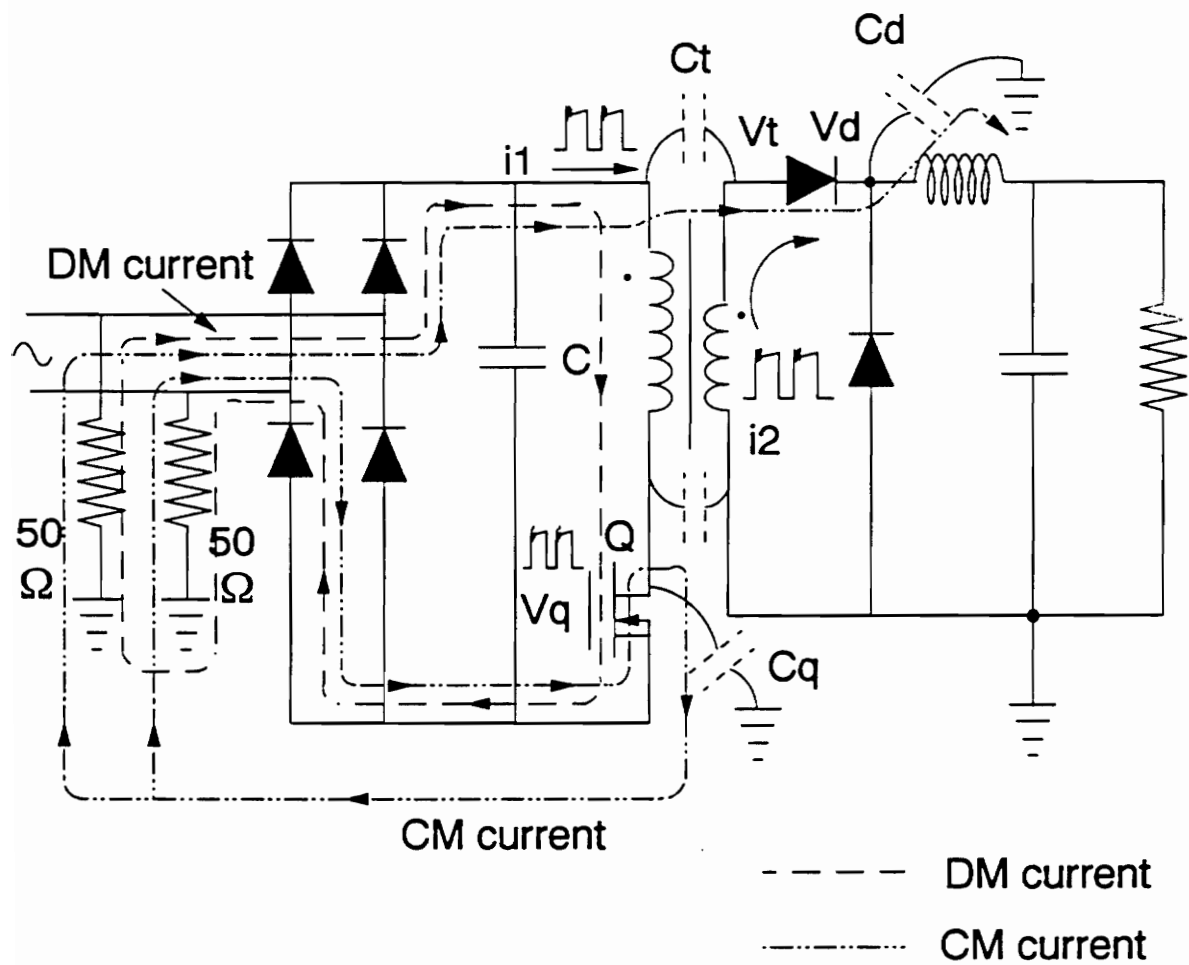


Figure 3.2.1 Conducted Noise Emission Sources In A Forward Converter Circuit

the pulsating current  $i_1$  is shunted by the bulk capacitor  $C$ , and the rest flows through the two  $50\ \Omega$  resistors via the rectifying diodes. If the impedance of the capacitor  $C$  is very small compared to  $100\ \Omega (= 50\ \Omega + 50\ \Omega)$ , then the D.M. noise coupling is small. The parasitic inductance of  $C$  must be kept low to avoid D.M. coupling at very high frequency.

The differential mode noise can, additionally, be generated by the rectifier diode and the freewheeling diode. This problem takes place due to the recovery characteristic of the diode. The recovery problems cause the voltage and the current spike in the secondary of the transformer. The spike couples into the primary of the transformer, as the differential mode noise.

### **3.3 Conducted C.M. Noise In A Switching Power Supply**

C.M. noise is caused by the current coupled through parasitic capacitors  $C_d$ ,  $C_q$  and  $C_t$ .  $C_d$ , and  $C_q$  are the parasitic capacitances between semiconductor ( diode or transistor ) and chassis, and  $C_t$  is the transformer inter-winding capacitance. The voltages across the three parasitic capacitors are high-voltage high-frequency waveforms. Displacement current generated by these voltage waveforms flows through the  $50\ \Omega$  resistors and causes the measured EMI emission. If the heat sinks used for the semiconductor devices are physically tied to the chassis, then the values of  $C_q$  and  $C_d$  will be increased and so will the measured EMI. If the winding arrangement of the transformer is altered,  $C_t$  will be affected, and so will the measured conducted EMI. Since the C.M. noise is coupled through the parasitic capacitances, circuit layout and packaging make differences in the overall noise performance.

### 3.4 Radiated Noise

Besides the conducted EMI noise, there is radiated noise emission in the switching mode power supply. The higher the switching frequency is, the more efficient the radiated emission is. Similar to the conducted EMI noise, the radiated noise emission exists in two modes: the differential mode and the common mode.

#### 3.4.1 Differential Mode

D.M. radiated emission is caused by current loop. The strength of the field is related to the loop current  $I$ , the switching frequency  $f$  and the loop area  $A$ . The field strength is expressed as

$$E_{\max} = 263 \times 10^{-16} (f^2 A I) \frac{1}{r},$$

E: V/Meter  
A:  $\text{cm}^2$   
f: Hz  
I: A  
r: Meter

where  $E_{\max}$  is the strength of the field,  $f$  is the noise frequency,  $A$  is the area of the current loop, and  $r$  is the distance between the noise source and the receiver.

From Figure 3.4.1, it can be seen that the primary current  $i_1$  and secondary current  $i_2$  both contribute to radiation emission. In an off-line converter for low output-voltage application,  $i_2$  is much larger and therefore could be the main contribution to D.M. radiation. High frequency ringing is especially effective in radiating the noise as expressed by

the formula. Therefore, it is important to reduce transformer secondary current ringing to reduce the D.M. radiation. It is also important to minimize the loop current area in the layout of the circuit.

### **3.4.2 Common Mode**

The other way that the noise emission occurs is through the rod antenna effect. The electric field radiated by a rod antenna is expressed as follow [6]:

$$E_{\max} = 12.6 \times 10^{-7} (fI) \frac{1}{r},$$

E: V/Meter  
f: Hz  
l: Meter  
I: A  
r: Meter

where  $E_{\max}$  ( V/m ) is the strength of the field,  $f$  ( Hz ) is the noise frequency,  $l$  ( m ) is the length of the rod,  $I$  ( A ) is the current in the rod, and  $r$  ( m ) is the distance between the rod and the measurement equipment.

In a forward converter, a floating heat sink can act as a rod antenna. The longer the antenna length, the larger the emission.

Typically, in the conducted EMI, the differential mode noise is dominant in the lower frequency range around the switching frequency. The reason is that the parasitic capacitance is generally very small, so the capacitive impedance is very high. The high impedance blocks the common mode current path, the common mode noise current is

very small comparing to the differential mode noise current. Within the high frequency range, in view of that the differential mode current comes from the pulsating current, the fundamental components of the current is the highest noise in the spectrum of the pulsating current. In the higher frequency range, the amplitude of the harmonics of the pulsating current is insignificant, then it is negligible. But the ringing current, with very high frequency, can go through the parasitic capacitance easily. So, in the high frequency range, the common mode current is dominant.

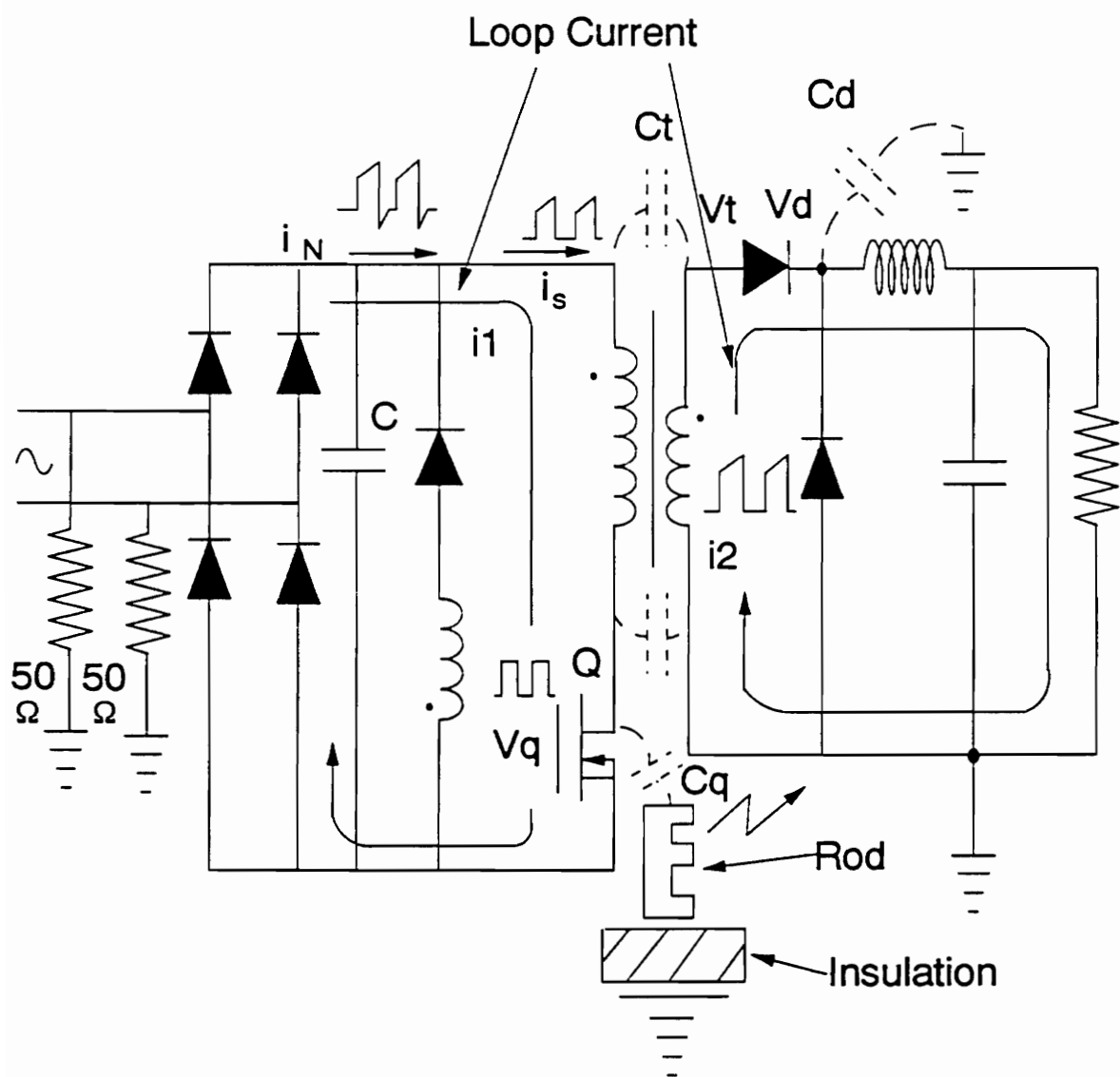


Figure 3.4.1 Source Of Radiated EMI Noise

## Chapter IV

### EFFECT OF FREQUENCY MODULATION ON NOISE SPECTRUM

From the discussion given in the last chapter, it is clear that the measured emission is periodical with respect to switching frequency. The emission, therefore, centers at the switching frequency and the harmonic frequencies. Concentration of emission power at these discrete frequencies makes it harder to meet EMI regulations. By modulating the switching frequency, side-bands are created, and the emission spectrum is smeared. The power is shattered into smaller pieces scattered around many side-band frequencies. This effect makes it easier for the resultant emission spectrum to pass EMI regulations. The details are explained below.

In this chapter, the effect of frequency modulation on a noise spectrum is discussed. Mathematical results are described, and backed up by experimental results.

#### 4.1 Spectrum Of Single-Tone Modulation

If the waveform of the modulating signal is sinusoidal, it is single-tone modulation. Assuming the initial phase  $\phi$  is zero, and the signal  $S(t)$  is cosine wave form shown in Figure 4.1.

$$S(t) = A_c \cos \omega_c t, \quad (4.1)$$

where  $A_c$  is the amplitude,  $\omega = 2\pi f_c$ ,  $f_c$  is the carrier frequency.

The modulating signal is a single-tone and can be expressed as follow:

$$\Delta\omega(t) = \Delta\Omega \cos \omega_m t , \quad (4.2)$$

where  $\Delta\Omega$  is the amplitude of the modulating signal,  $\omega_m$  is the angle frequency of the modulating signal. Substituting Equation ( 4.2 ) into Equation ( 4.1 ) leads to the expression of the modulated signal.

$$\begin{aligned} S_m(t) &= A_c \cos(\omega_c + \Delta\omega)t \\ &= A_c \cos(\omega_c + \Delta\Omega \cos \omega_m t)t \\ &= A_c \cos \omega_i(t)t , \end{aligned}$$

where  $\omega_i(t) = \omega_c + \Delta\Omega \cos \omega_m t$  .

Simplifying the equation,

$$\theta(t) = \int_0^t \omega_i(t)dt = \omega_c t + \frac{\Delta\Omega}{\omega_m} \sin \omega_m t + \theta_0.$$

Since the initial phase is zero, that is,  $\theta_0 = 0$ ,

$$\theta(t) = \omega_c t + \beta \sin \omega_m t,$$

where  $\beta = \Delta\Omega/\omega_m = \Delta f/f_m$ , and is called modulation index. The effect of modulating index on spectrum is discussed later.

So

$$S_m(t) = A_c \cos(\omega_c t + \beta \sin \omega_m t)$$

$$\begin{aligned}
&= A_c \cos \omega_c t \left[ J_0(\beta) + 2 \sum_{\lambda=1}^{\infty} J_{2\lambda}(\beta) \cos 2\lambda \omega_m t \right] \\
&\quad - A_c \sin \omega_c t \left[ 2 \sum_{\nu=1}^{\infty} J_{2\nu-1}(\beta) \sin(2\nu-1) \omega_m t \right] \\
&= A_c \left[ \cos \omega_c t \sum_{n=-\infty}^{\infty} J_n(\beta) \cos n \omega_m t - \sin \omega_c t \sum_{n=-\infty}^{\infty} J_n(\beta) \sin n \omega_m t \right],
\end{aligned}$$

where  $J_n(\beta) = \frac{1}{2\pi} \int_{-\pi}^{\pi} e^{j(\beta \sin \theta - n\theta)} d\theta$ , it is first order Bessel Function.

Because  $J_n(\beta) = (-1)^n J_{-n}(\beta)$ ,

$$S_m(t) = A_c \sum_{n=-\infty}^{\infty} J_n(\beta) \cos(\omega_c + n \omega_m)t. \quad (4.3)$$

Figure 4.1 shows the effect of frequency modulation on the frequency spectrum, where  $f_c$  is the carrier frequency,  $f_m$  is the modulating frequency, and  $\Delta f$  is the amplitude of frequency change. As can be seen from Figure 4.1 ( b ), side-band harmonics are generated, and the magnitude at  $f = f_c$  is reduced compared to the unmodulated signal ( i.e.  $A_1 < A$  ). It looks as if the bar in Figure 4.1 ( a ) is shattered and scattered around the side-band frequencies shown in Figure 4.1 ( b ). Notice that the frequency difference between each two adjacent side-band harmonics is  $f_m$ .

## 4.2 $\beta$ Value

The extent of shattering and the magnitude of the resultant spectrum depend on the modulation index  $\beta \equiv \Delta f / f_m$  [3]. Figure 4.2.1 shows the actual spectra for several  $\beta$  values for single-tone modulated signal. The larger the  $\beta$  value, the more evenly distributed is the spectrum. It can be seen from Figure 4.2.1 that when  $\beta = 5$ , the maximum amplitude of the harmonic is about one third of the unmodulated signal amplitude, and is one fourth of it when  $\beta = 10$ .

It should be noticed, however, the results shown in Figure 4.2.1 are purely mathematical. When measured from a spectrum analyzer, the results are affected by the resolution bandwidth ( RBW ) setting of the analyzer, as described in Section 2.3.1. If the RBW is set smaller than  $f_m$  then measurement result is closed to the mathematical result. Figure 4.2.2 shows the measurement result which verifies the mathematical result shown in Figure 4.2.1.

## 4.3 Carson's Rule

Carson's Rule states two important characteristics of an FM signal [3]:

( a ) Total power of a signal is unaffected by the frequency modulation. The total power of a signal is equal to the summation of the square of each harmonic amplitude. Referring to Figure 4.2.1, this means  $A^2 = A_1^2 + 2(A_2^2 + A_3^2 + \dots)$

( b ) 98% of the total power of a frequency-modulated signal is contained inside the bandwidth  $B_T$ , where  $B_T = 2 ( \beta + 1 ) f_m$ . Practically speaking, this means the side-band harmonic frequency ranges from  $( f_c - B_T / 2 )$  to  $( f_c + B_T / 2 )$ . Using the definition of  $\beta = \Delta f / f_m$ ,  $B_T = 2 \Delta f ( \beta + 1 ) / \beta$ . If  $\beta \gg 1$ , then  $\beta_T \approx 2 \Delta f$ .

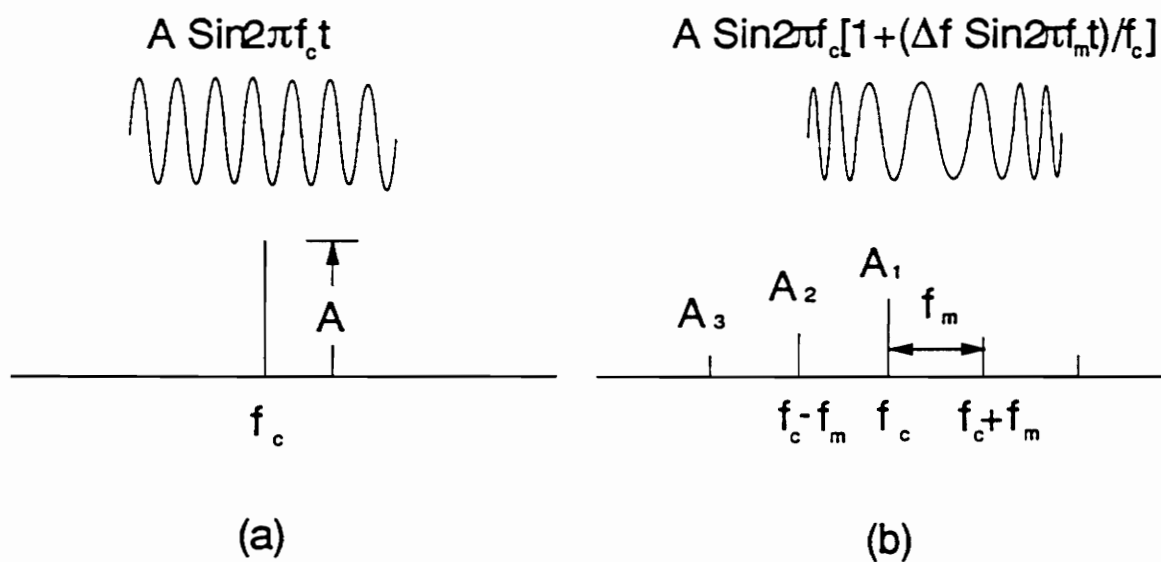


Figure 4.1 Modulated Sinusoidal Waveform

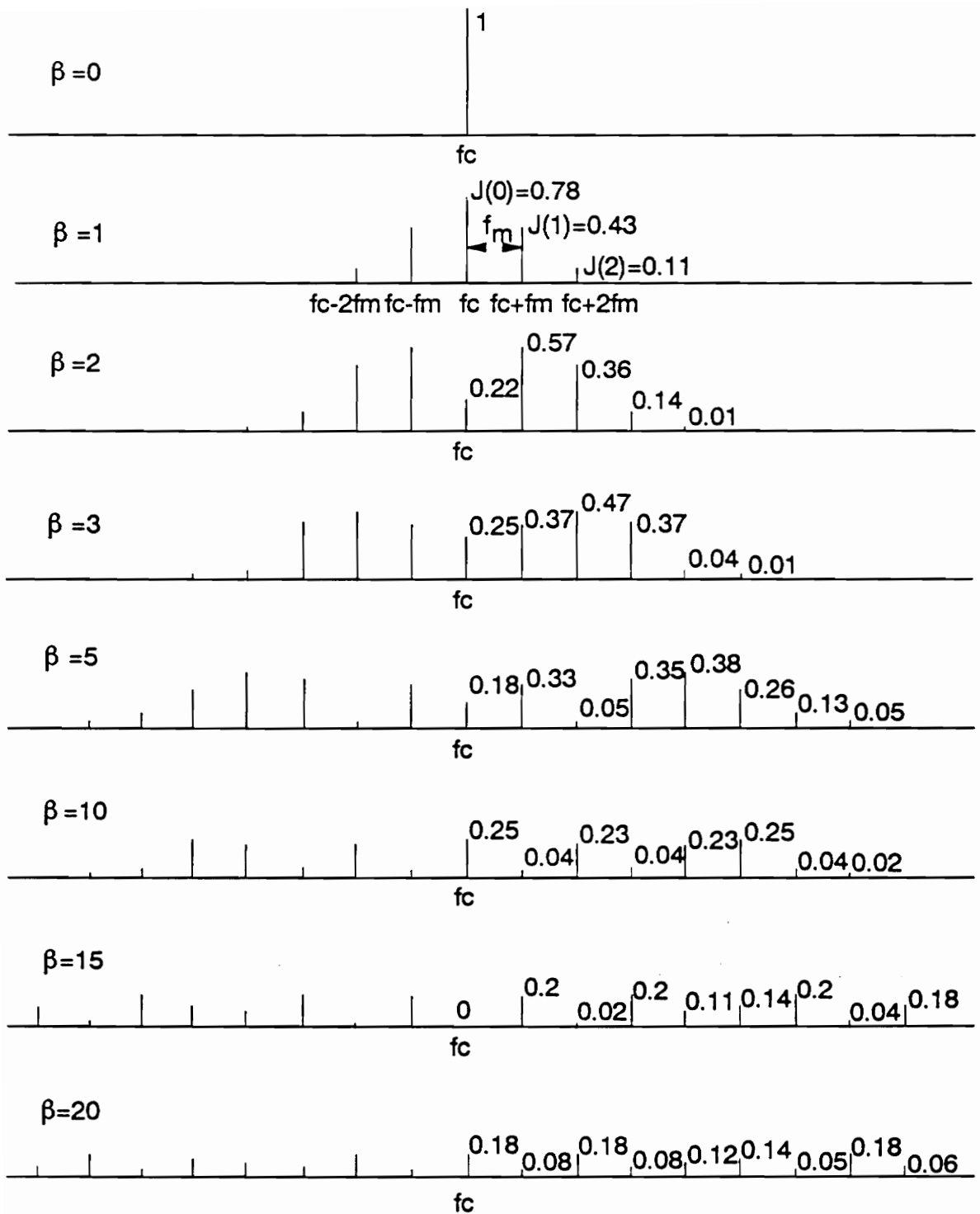


Figure 4.2.1 Effect Of  $\beta$  -Value On The Spectrum Of A Single-Tone Modulated Sinusoidal

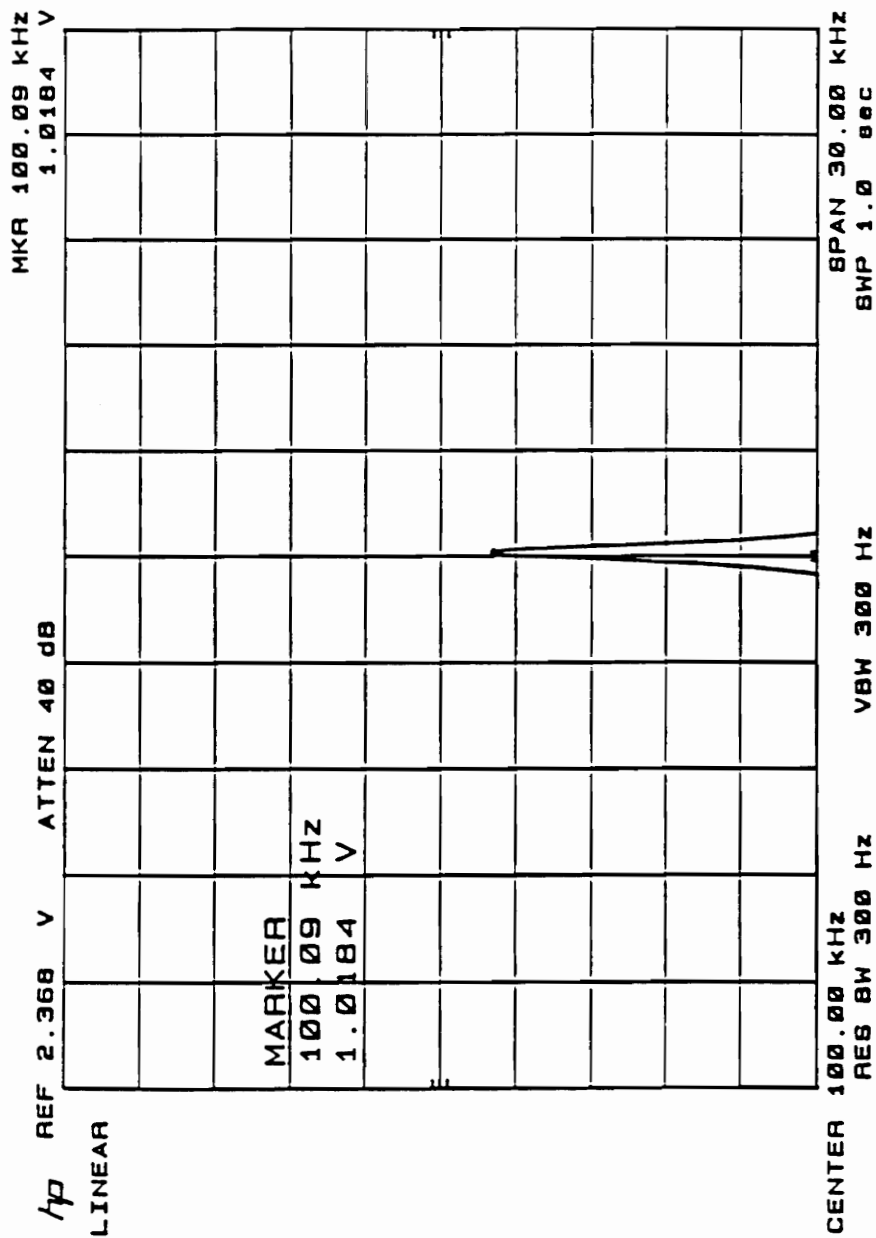
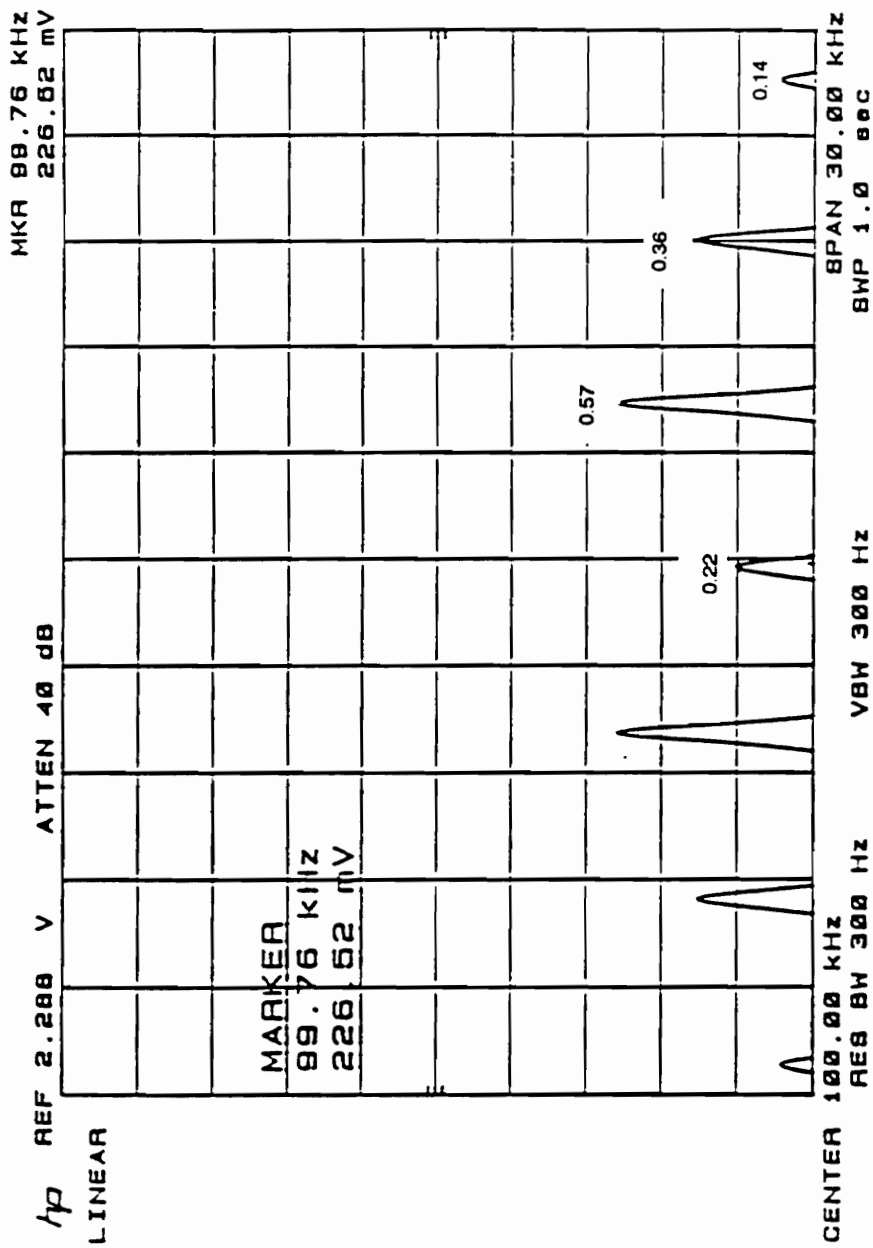
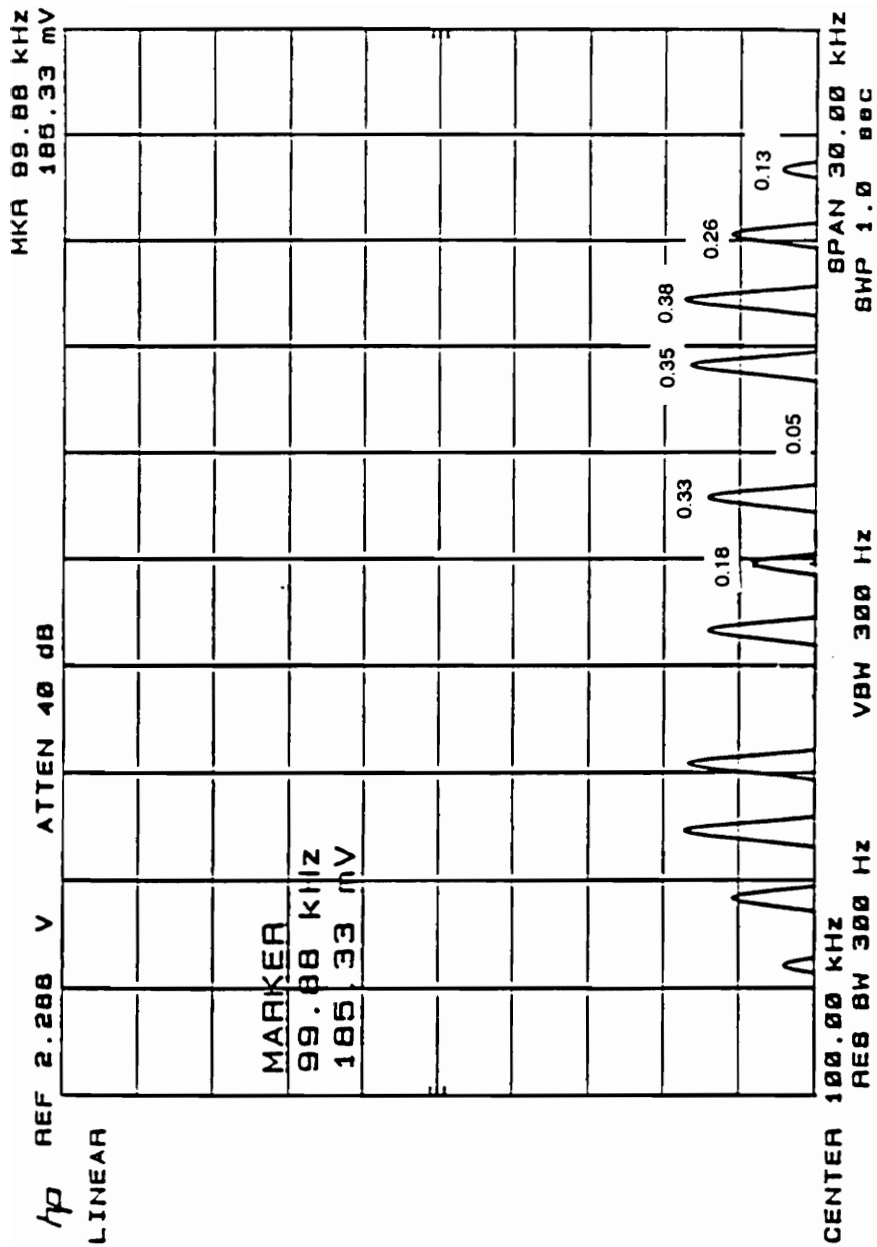


Figure 4.2.2 (a) Unmodulated Signal,  $f_c = 100$  kHz, RBW = 300 Hz



$\Delta f = 10 \text{ kHz}$ ,  $f_m = 5 \text{ kHz}$ ,  $\beta = 2$ ,  $\text{RBW} = 300 \text{ Hz}$

Figure 4.2.2 (b) Spectra Of A FM-Modulated Signal



$\Delta f = 10 \text{ kHz}$ ,  $f_m = 2 \text{ kHz}$ ,  $\beta = 5$ ,  $\text{RBW} = 300 \text{ Hz}$

Figure 4.2.2 (c) Spectra Of A FM-Modulated Signal

For example,

if

$$\Delta f = 15 \text{ kHz},$$

$$f_m = 400 \text{ Hz},$$

then

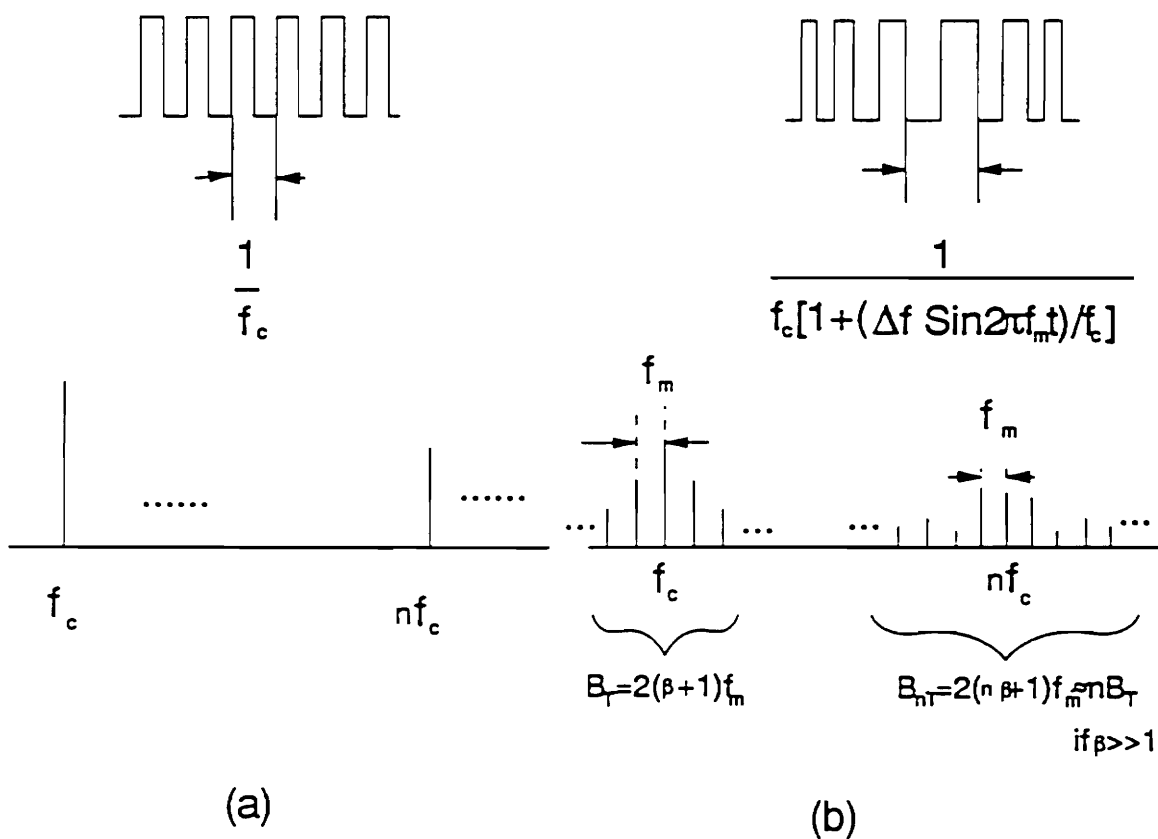
$$\beta = 39,$$

$$B_T = 2(39+1)400 = 32 \text{ kHz}.$$

This result has practical ramification in using the frequency-modulation scheme to reduce EMI noise, because in some application, the design should avoid certain forbidden frequency band.

#### 4.4 Spectrum Of Pulse Train

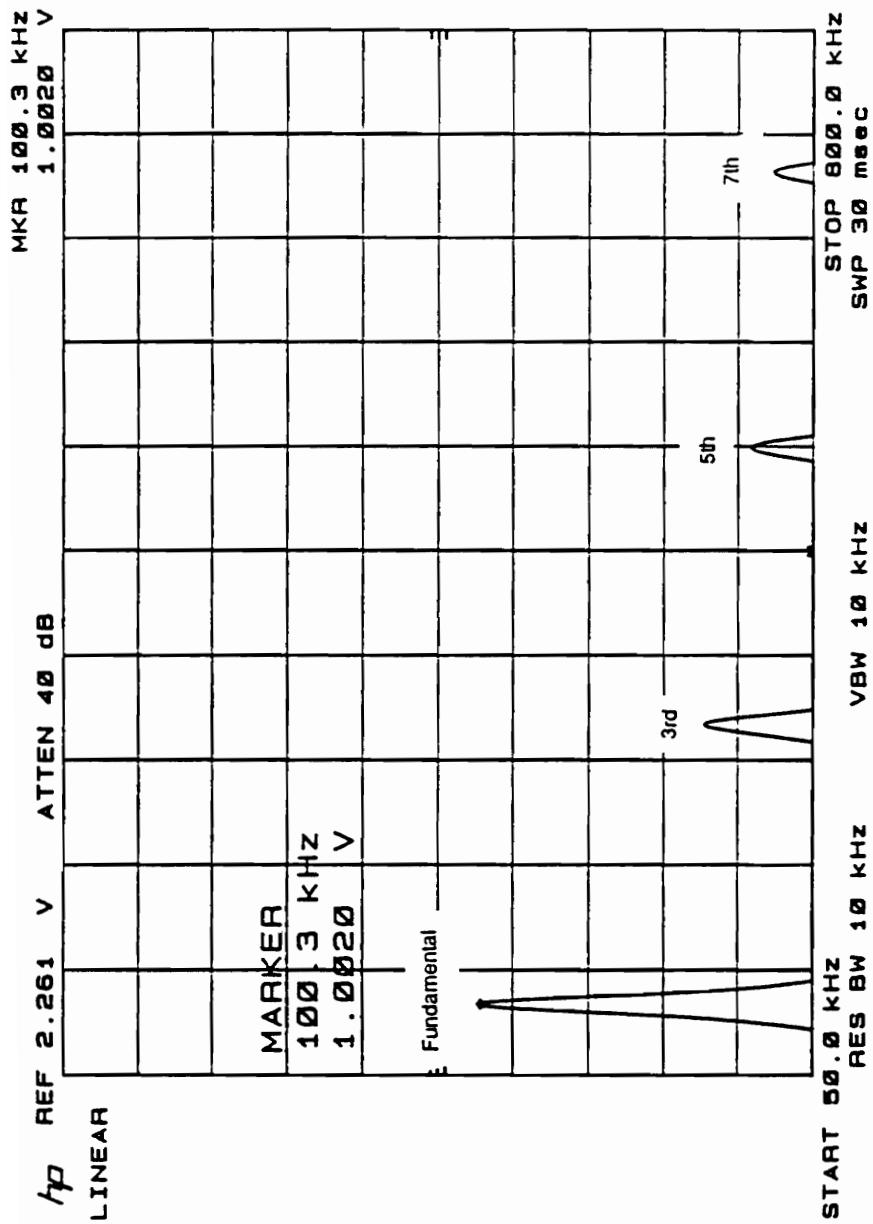
If the unmodulated waveform is a pulse train, like the waveforms in many power supply circuits, then it itself contains harmonics. To avoid confusion, these harmonics are referred to as switching-frequency harmonics. Frequency modulation of the pulse-train waveform shatters each of the switching harmonic components into side-band harmonics as shown in Figure 4.4.1. Notice, however, that the frequency difference between each



(a) Square Waveform

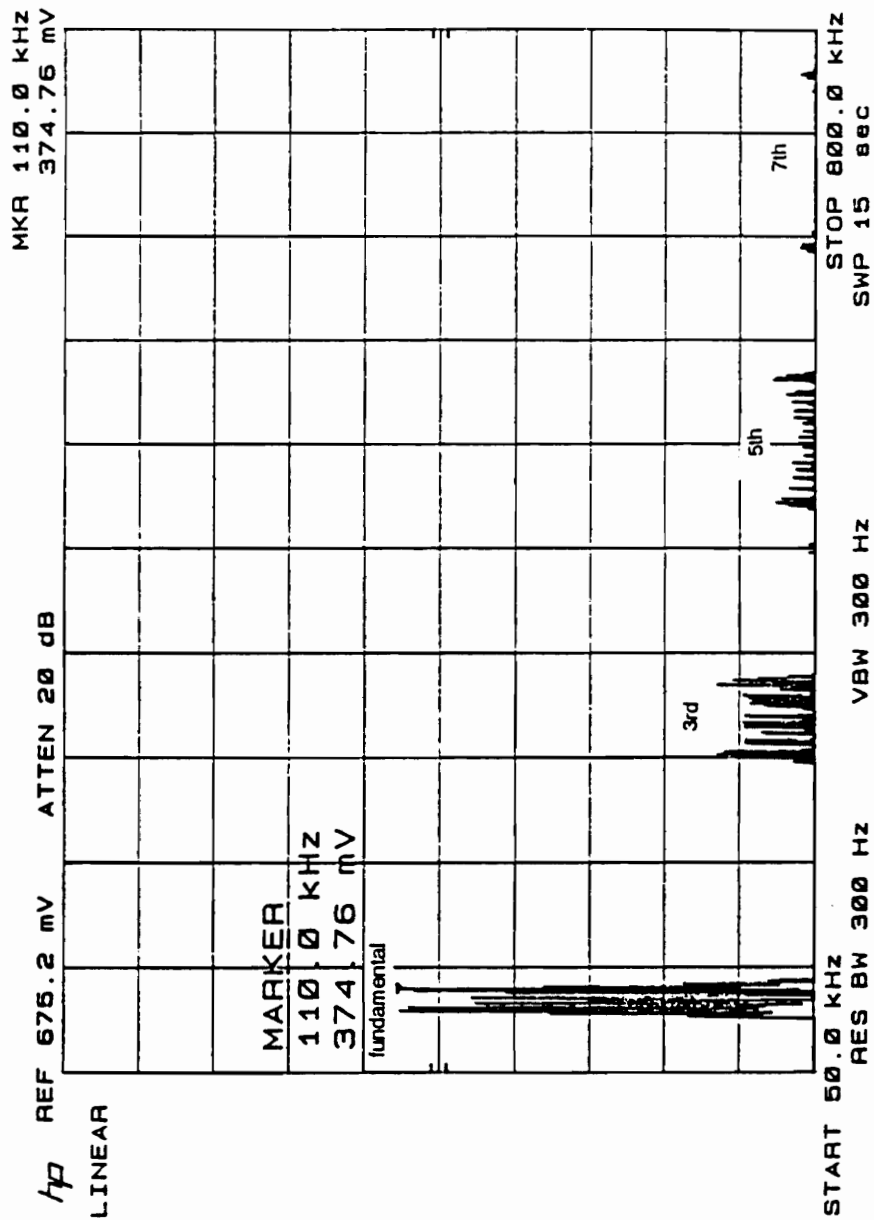
(b) Frequency-Modulated Square Waveform

Figure 4.4.1 Spcetrum Of Modulated Pulse Train



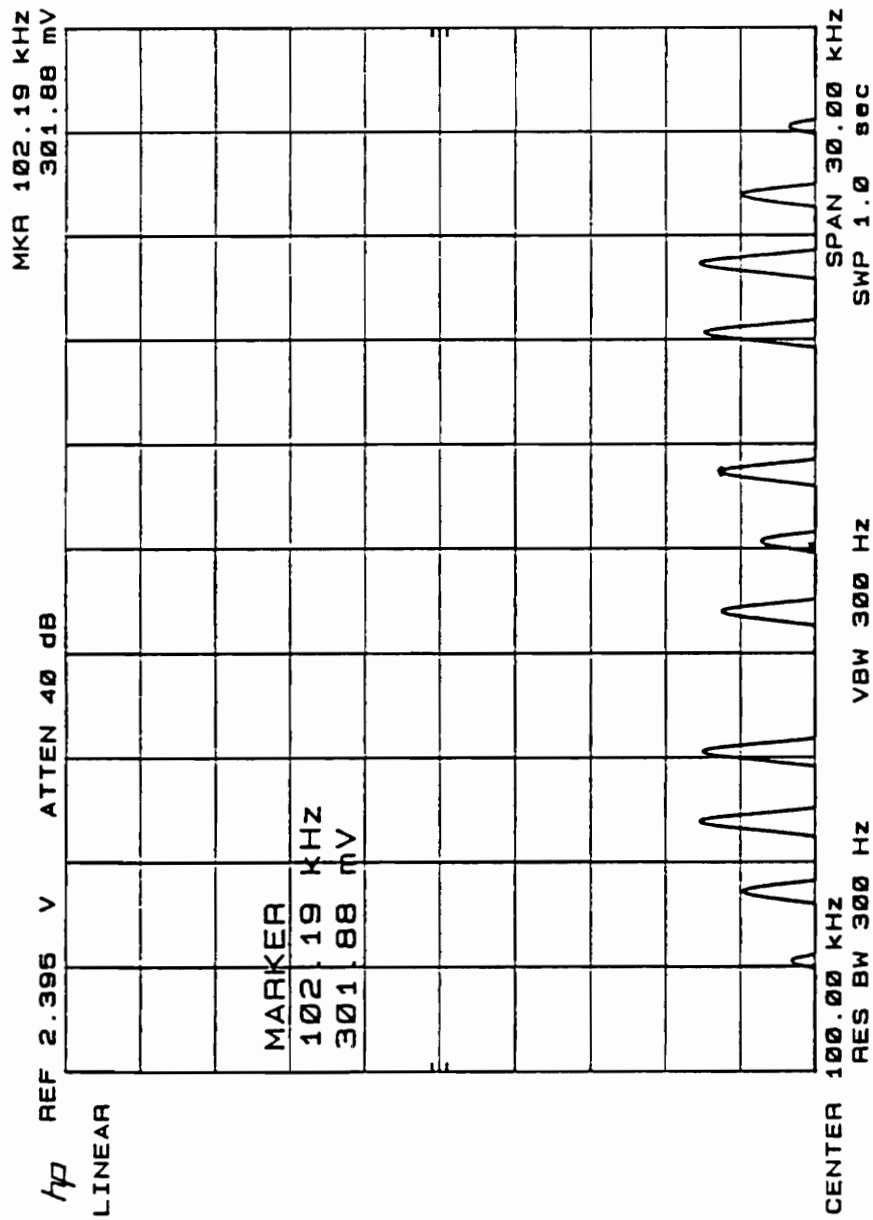
$f_c = 100 \text{ kHz}$ ,  $\text{RBW} = 10 \text{ kHz}$

Figure 4.4.2 (a) Spectra Of An Unmodulated Pulse-Train Signal



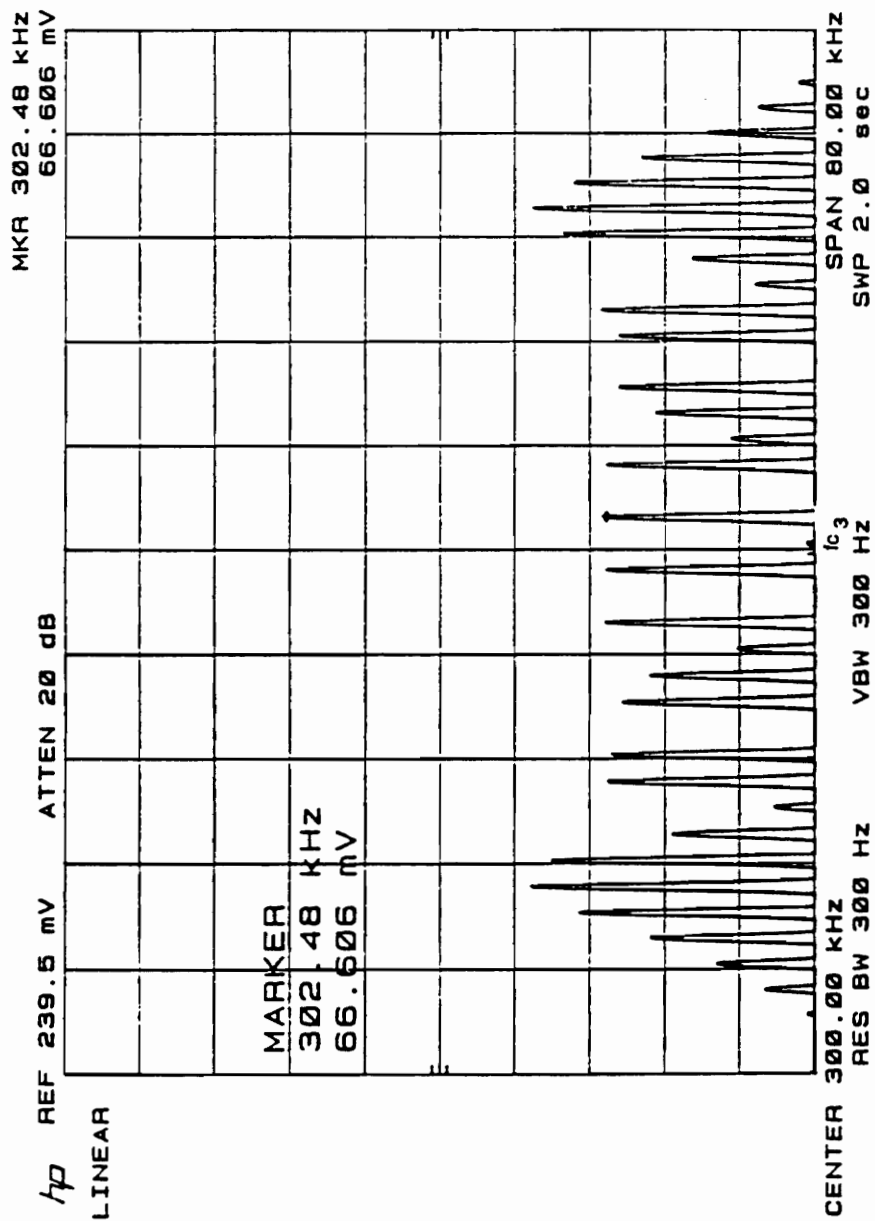
$f_c = 100 \text{ kHz}$ ,  $f_m = 2 \text{ kHz}$ ,  $\Delta f = 10 \text{ kHz}$ ,  $\beta = 5$ ,  $\text{RBW} = 10 \text{ kHz}$

Figure 4.4.2 (b) Spectrum Of A Modulated Pulse-Train Signal



$\Delta f = 10 \text{ kHz}$ ,  $f_m = 2 \text{ kHz}$ ,  $\beta = 5$ ,  $\text{RBW} = 300 \text{ Hz}$

Figure 4.4.2 (c) Side-Band Harmonics Of The Fundamental Switching Harmonic :



$$\Delta f_3 = 30 \text{ kHz}, f_m = 2 \text{ kHz}, \beta = 15, \text{RBW} = 300 \text{ Hz}$$

Figure 4.4.2 (d) Side-Band Harmonics Of The Third Switching Harmonic

two adjacent side-band harmonics is still  $f_m$ , but the modulation index of each switching harmonic  $\beta_n$  is different, and  $\beta_n = n\beta$ , where  $n$  is the number of switching harmonics of the unmodulated pulse train ( e.g.  $n = 2$  for second switching-frequency harmonics ). Since the modulation index changes with each switching harmonic, the scattering effect of each switching harmonics is different. For example, if  $\beta = 5$ , then the third switching harmonic is scattered with modulation index of 15. Carson's rule applies to each harmonic, i.e.,  $B_{nT} = 2(n\beta + 1)f_m$ . If  $\beta \gg 1$ , then  $B_{nT} = nB_T$ . This implies that the bandwidth of side-band harmonics increases with  $n$ , the number of switching harmonic. It is possible that the side-band harmonics of different switching-frequency harmonics overlap. The general tendency of frequency modulation is to spread out the energy of each switching harmonic. The higher the harmonic number, the more even is the spread-out energy. Figure 4.4.2 shows spectrum of pulse-train signals. Figure 4.4.2 ( a ) shows the spectra of an unmodulated pulse train. Figure 4.4.2 ( b ) shows the spectra when the signal in Figure 4.4.2 ( a ) is modulated by a 2 kHz modulating signal, and  $\Delta f$  is 100 kHz,  $\beta$  is 5 for fundamental in this case. Side-band harmonics generation is evidently shown. It is also clearly shown that the bandwidth of the side-band harmonics increases with  $n$  as discussed above. Figure 4.4.2 ( c ) shows the detail of the side-band harmonics of the fundamental switching harmonic. Figure 4.4.2 ( d ) shows the detail of the side-band harmonics of the third switching harmonic. As described above, for the third switching harmonic,  $\beta$  value is  $3 \times 5 = 15$ . This experimental result agrees with the theoretical result of  $\beta = 15$  shown in Figure 4.2.1.

It is described in Chapter III that the noise generated by the power supply, either D.M. noise or C.M. noise, is a pulse-train waveform. The effect of FM on the power supply noise spectrum is similar to that depicted by Figure 4.4.1.

## **Chapter V**

### **VERIFICATION OF POWER SUPPLY NOISE**

A frequency-modulated forward converter is designed and built to test the theory described in Chapter IV. The effect of FM on noise performance will be demonstrated. A variety of parameters involved in EMI reduction will be pointed out.

The theoretical spectrum analysis discussed in Chapter IV is purely mathematical. The noise source waveforms of an off-line power supply are not exactly a pulse train as discussed in Section 4.4. When a modulation is introduced into the PWM frequency, the noise source waveforms amplitude may be affected, and the analytical results given in Section 4.4 is not exactly correct. However, if the modulation frequency range  $\Delta f$  is much less than the carrier frequency  $f_c$ , then the results given in Section 4.4 is approximately correct. In practical implementation,  $\Delta f/f_c \leq 15\%$ .

In the practical EMI measurement, the results are affected by yet another constraint, that is the resolution bandwidth ( RBW ) required by the FCC and VDE rules. Section 5.2 gives examples of the results obtained by using FM. In addition, there are practical constraints which the designers have to consider. These constraints and the relationship of FM parameters are discussed in Section 5.3. Section 5.4 discusses possible adverse effects caused by FM.

#### **5.1 Description Of Testing Circuit**

In this project, an off-line forward converter is used to verify the theory. Figure 5.1 shows the overall circuit. The key operating parameters of the converter are listed below:

Switching Frequency: 90 kHz nominal

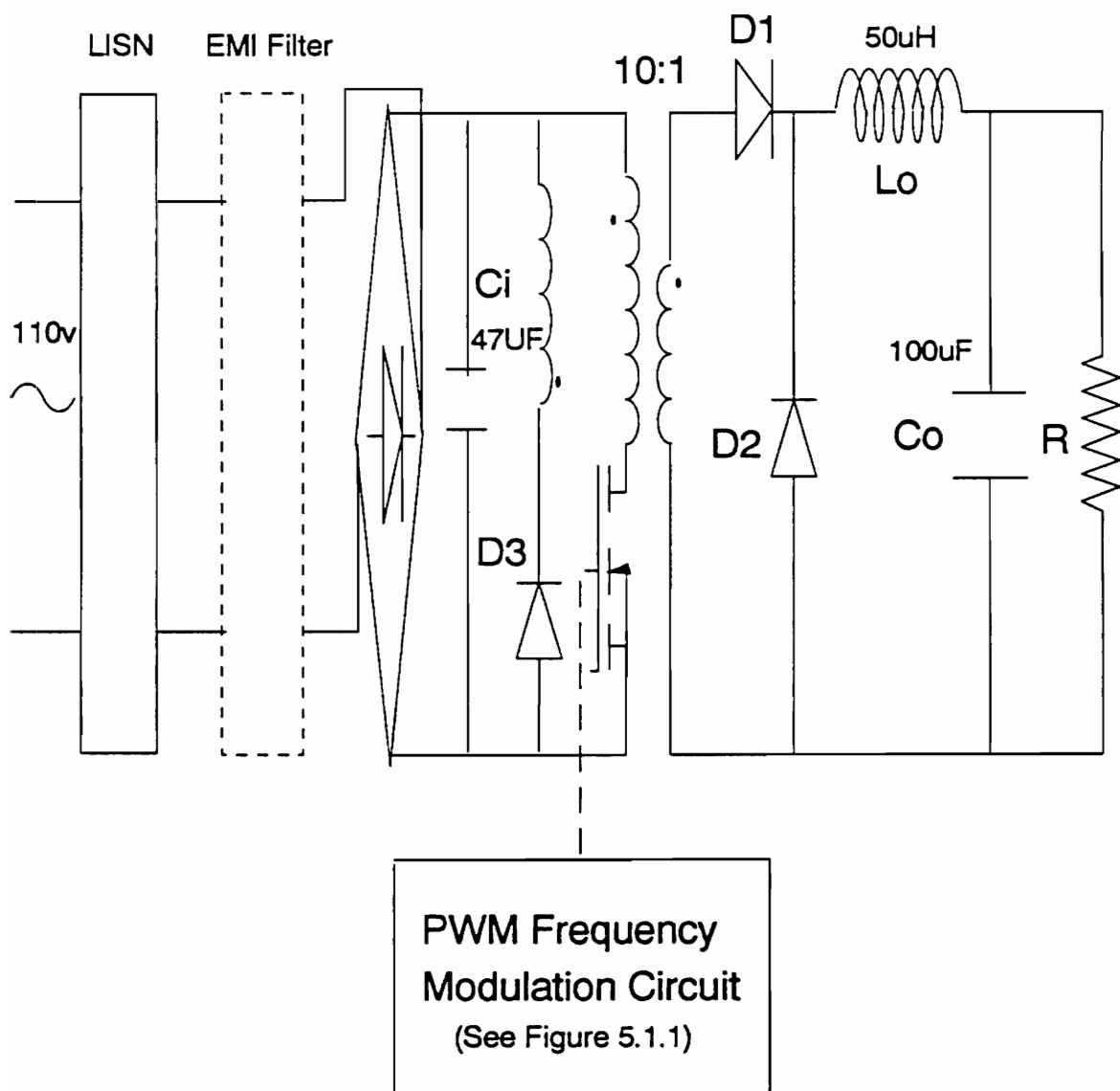


Figure 5.1 Testing Circuit

Modulating frequency (  $f_m$  ): adjustable

Frequency Variation (  $\Delta f$  ): adjustable but should not over 15 kHz

Output Voltage: 3 V

Output Current: 5 A

Input Line Voltage: 110 Vrms, 60 Hz

Control Strategy: Single output voltage feedback loop with crossover bandwidth of 3 kHz

### **5.1.1 PWM-Frequency Modulation Circuit**

Figure 5.1.1.1 shows the PWM-frequency modulation controller. The PWM frequency is modulated by the modulation signal  $V_c$  through Pin 5. The frequency is determined by RC time constant of  $C_t$  and (  $R_m // R_t$  ). The resistors and capacitors connected to Pin 1 and Pin 3 (  $R_1, R_2, R_3, C_1, C_2, C_3$  ) are for feedback compensation. Pin 14 is used for driving the main power switch of the converter.

As mentioned earlier, the PWM frequency is determined by the RC time constant of  $C_t$  and (  $R_m // R_t$  ). Variable  $R_m$  is achieved by modulation of a bipolar transistor through base signal as shown in Figure 5.1.1.2. Referring to 5.1.1.2,  $V_{ref}$  and  $R_c$  are used to set the DC operating point of the BJT, and the signal generator provide frequency modulation. In other words, the effective resistance between collector and emitter of the transistor 2N2369 provides the modulating effect on the PWM frequency.  $R_t$  is a manually adjustable resistance, and is used to give a central switching frequency  $f_c$ . Figure 5.1.1.3 shows the frequency variation due to control voltage  $V_c$ . It can be seen that, when  $V_c$  is lower than 0.6V or higher than 1.2V, the characteristic becomes nonlinear due to transistor cut off or saturation. When  $V_{be}$  is lower than 0.6V, the transistor is essentially

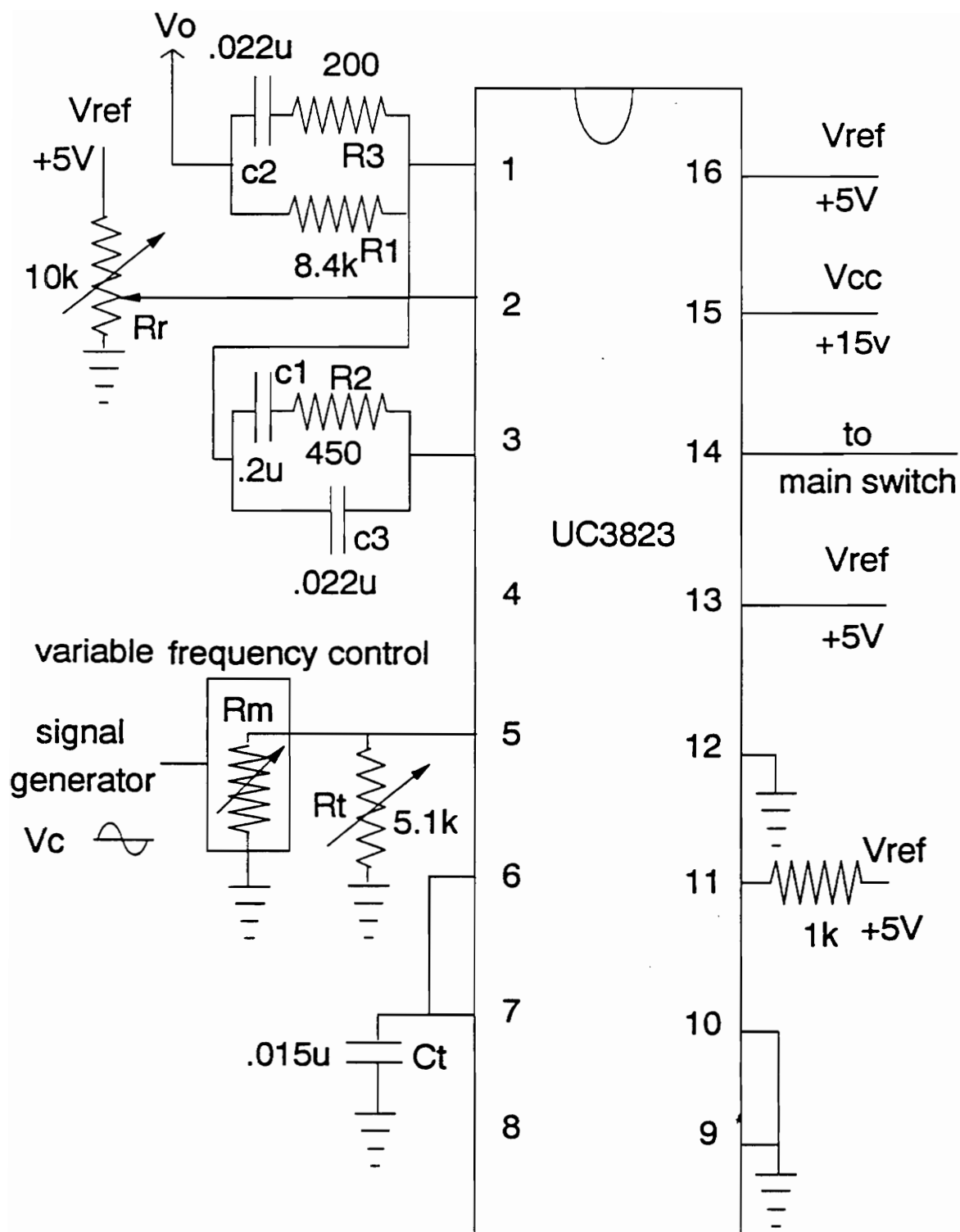


Figure 5.1.1.1 Frequency - Modulated PWM Controller

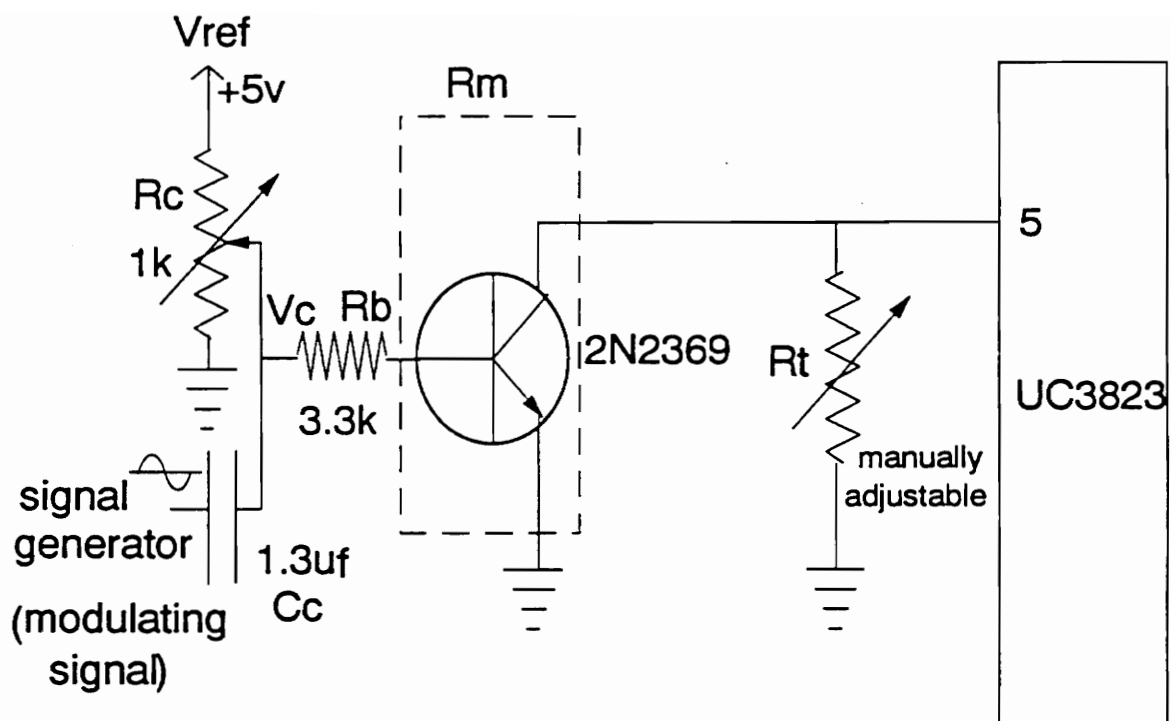


Figure 5.1.1.2 Variable Switching Frequency Circuit

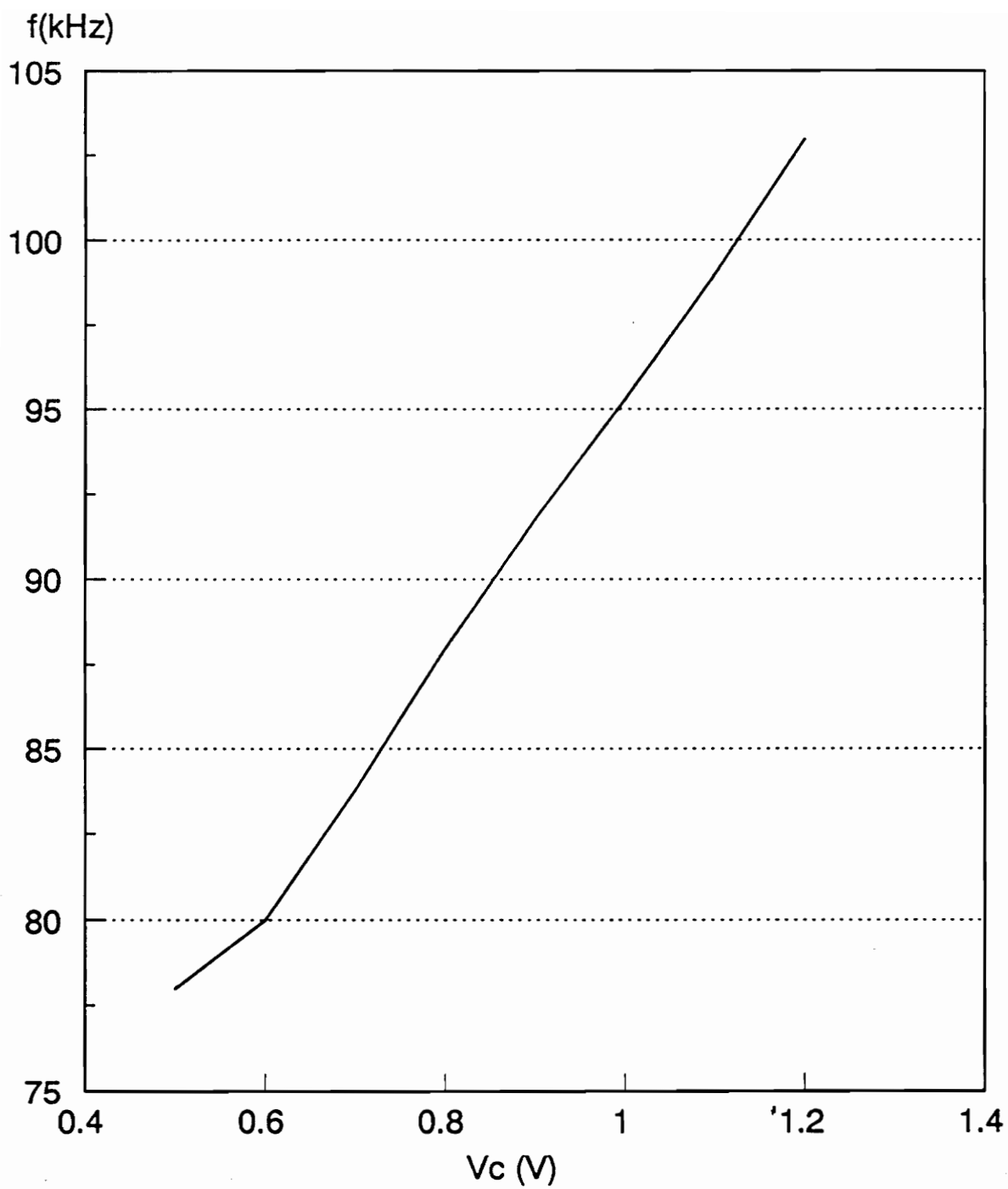


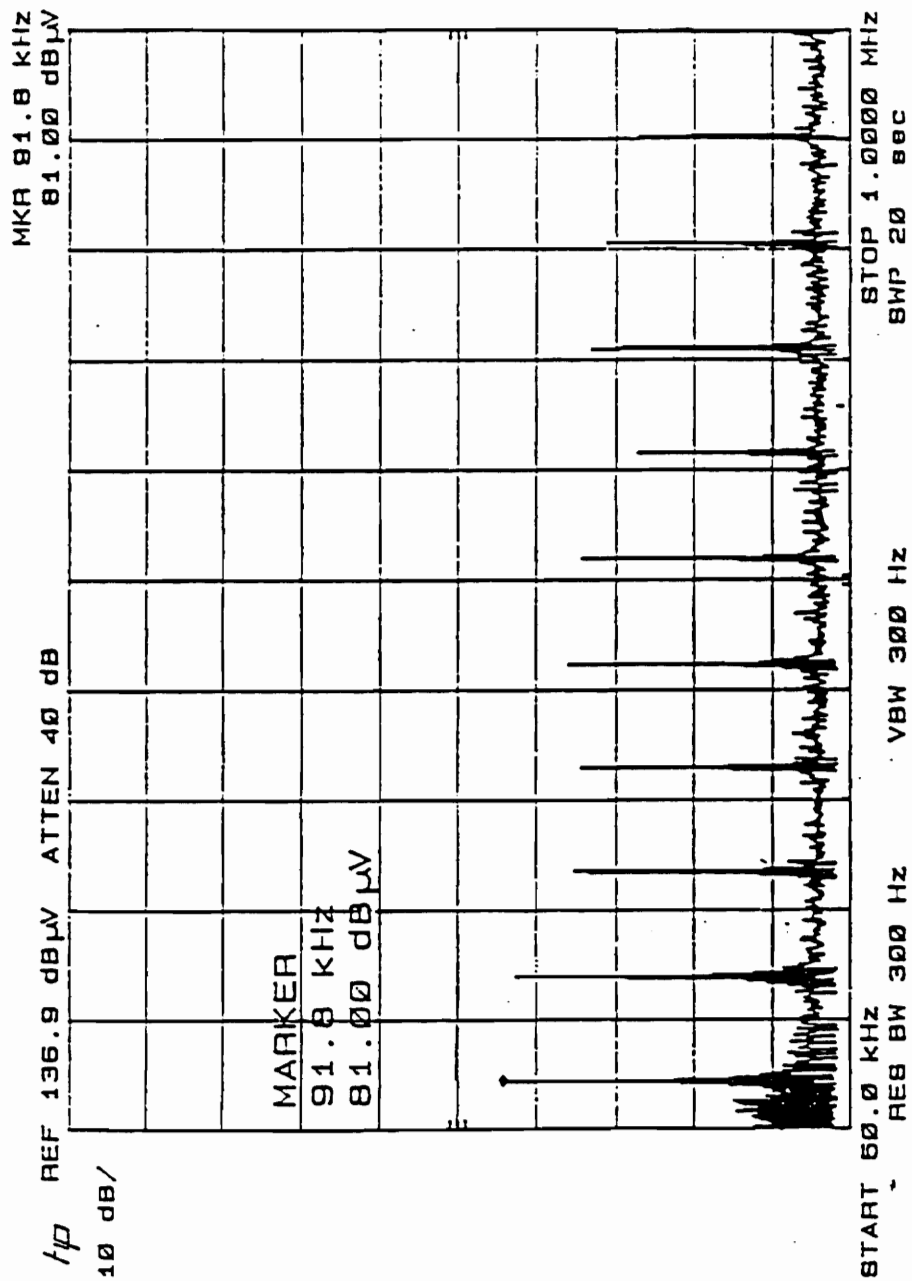
Figure 5.1.1.3 Frequency vs. Control Voltage Amplitude

open, and when  $V_{be}$  is higher than 1.2V, the transistor is essentially saturated. The central frequency ( i.e.  $f_c$  ), is determined by  $R_t$  and  $C_t$ .  $R_t$  can be adjusted to provide the desirable  $f_c$ . In Figure 5.1.1.3,  $f_c$  of 90 kHz was used.

## 5.2 Measurement Results

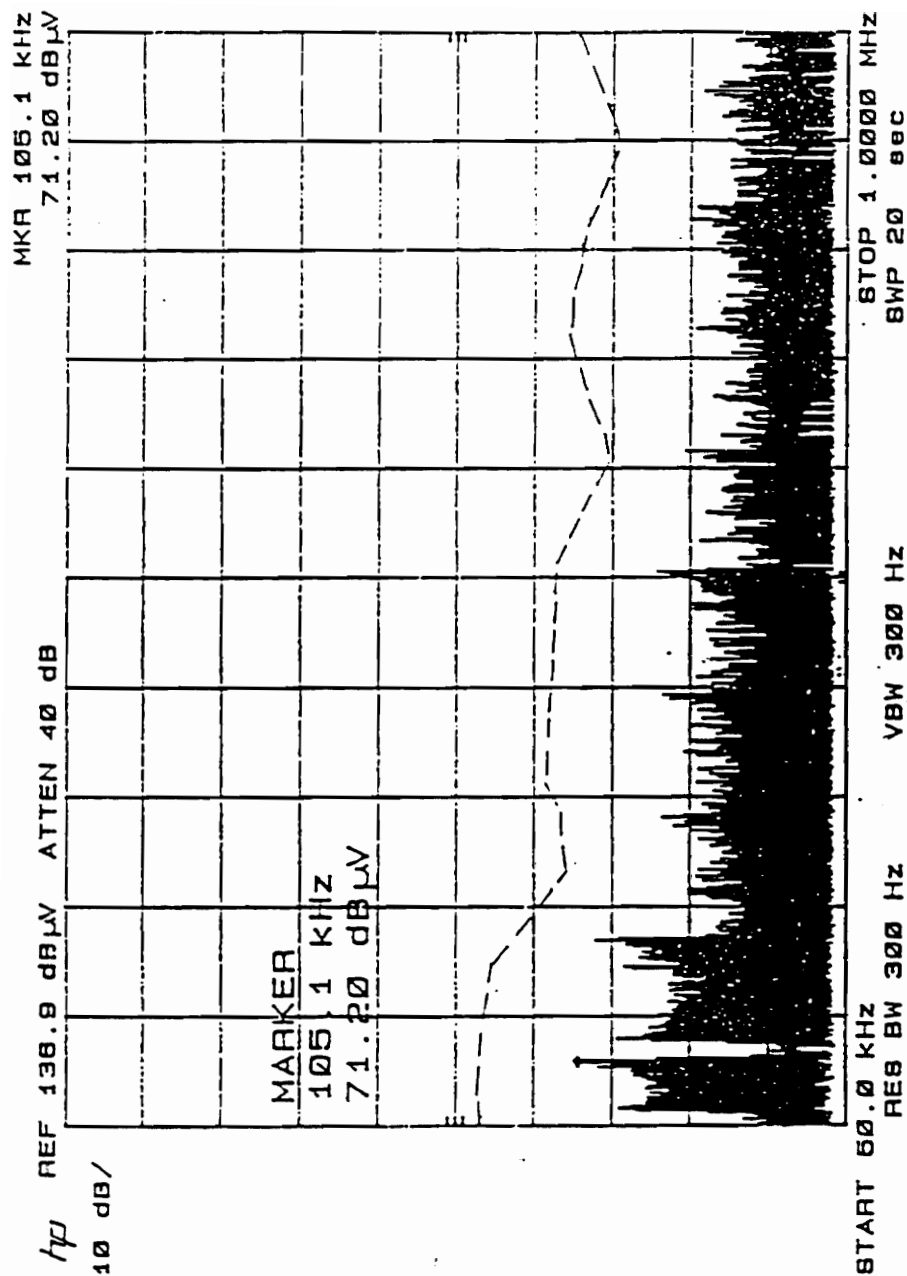
A Hewlett-Packard 8568B spectrum analyzer is used for EMI measurement. Peak detector is used for all the tests, since quasi-peak is not available from this model. Figure 5.2.1 ( a ) shows the measured conducted EMI of the forward converter described above. RBW is set at 300 Hz. It is to be noticed that the spectrum analyzer RBW is not set exactly according to FCC regulations as summarized in previous chapters. This is done purposely to illustrate higher order switching frequency harmonics clearly. For the unmodulated case, the amplitude of the harmonics is clearly shown in Figure 5.2.1 ( a ). Figure 5.2.1 ( b ) shows the spectrum for the case when the PWM frequency is modulated by a signal with  $f_m = 400$  Hz and  $\Delta f = 15$  kHz. It can be seen from Figure 5.2.1 ( b ) that the switching frequency harmonics are shattered by the modulation, and the overall spectrum improves. The dotted line indicates the envelope of the unmodulated spectrum shown in 5.2.1 ( a ). Significant reduction is achieved with modulation scheme. These results are closed to the theoretical spectrum predicted in Chapter IV, because RBW is set small compared to the switching frequency.

Figure 5.2.2 shows the measurement results of the same power supply under the same condition, but with proper RBW. According to FCC and VDE rules, RBW should be set at 200 Hz for  $f < 150$  kHz, and 9 kHz for  $150 \text{ kHz} < f < 30\text{MHz}$ . Figure 5.2.2 ( a ) shows the noise spectrum between 50 kHz and 150 kHz, and Figure 5.2.2 ( b ) shows the spectrum between 150 kHz and 30 MHz for the unmodulated case. Notice from the figure that RBWs of 300 Hz and 10 kHz were used ( instead of 200 Hz and 9 kHz as required ),



Unmodulated,  $f_c = 90 \text{ kHz}$ ,  $\text{RBW} = 300 \text{ Hz}$

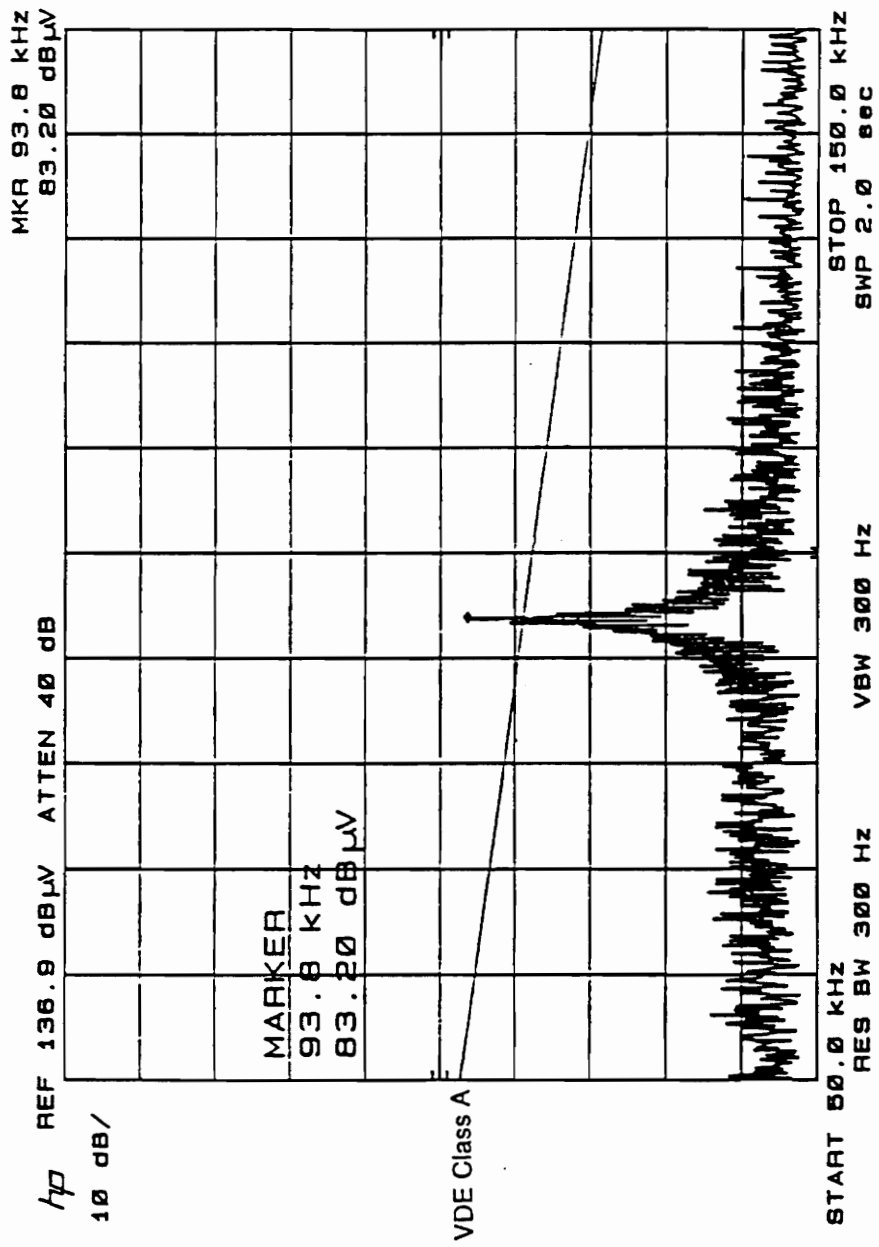
Figure 5.2.1 (a) Noise Spectrum Of A Forward Converter



Dotted line: envelop of unmodulated spectrum

Modulated,  $\Delta f = 15$  KHz,  $f_m = 400$  Hz, RBW = 300 Hz

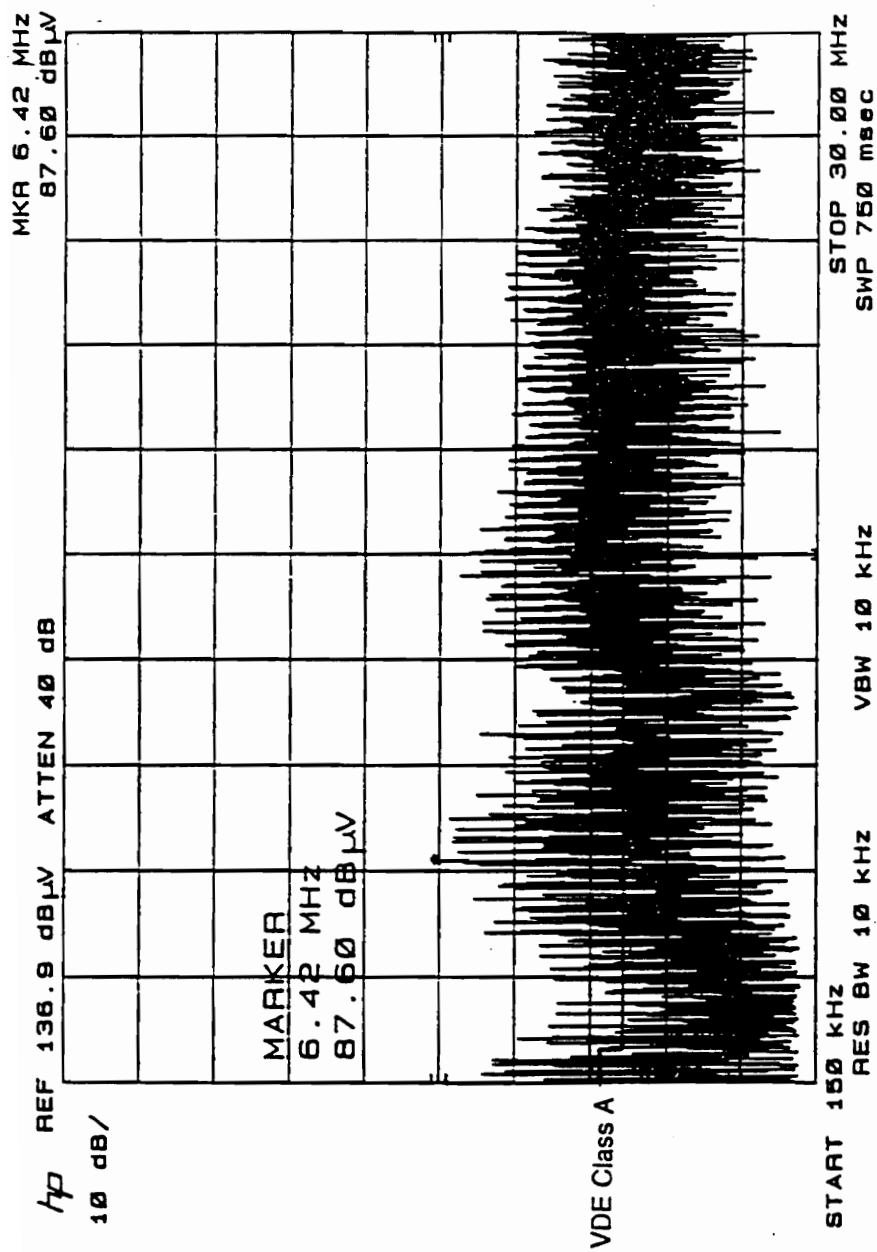
Figure 5.2.1 (b) Noise Spcctrum Of A Forward Converter



Solid line: VDE Class A

Unmodulated,  $f_c = 90\text{ kHz}$ ,  $50\text{ kHz} \sim 150\text{ kHz}$ ,  $\text{RBW} = 300\text{ Hz}$

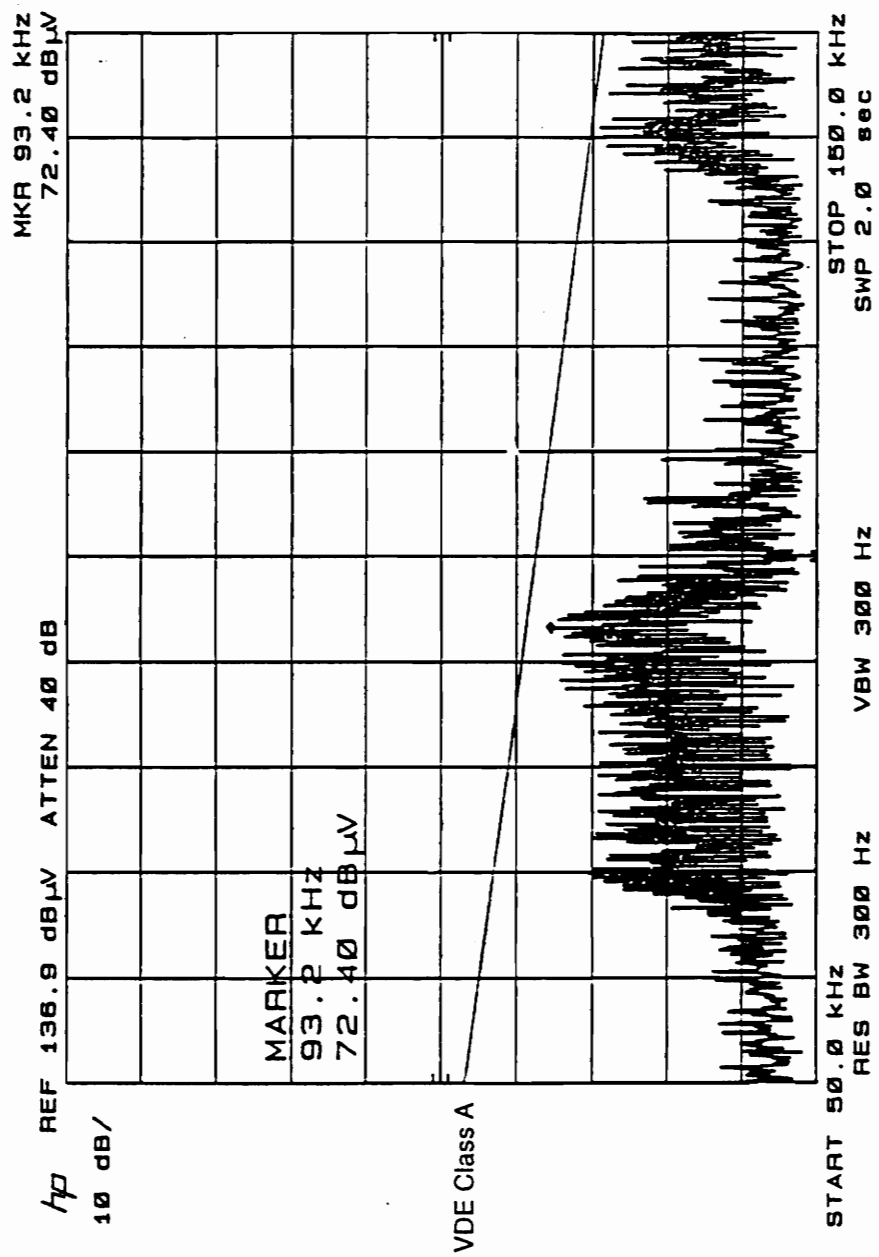
Figure 5.2.2 (a) Noise Spectrum Of A Forward Converter



Solid line: VDE Class A

Unmodulated,  $f_c = 90$  kHz, 150 kHz--30 MHz, RBW = 10 kHz

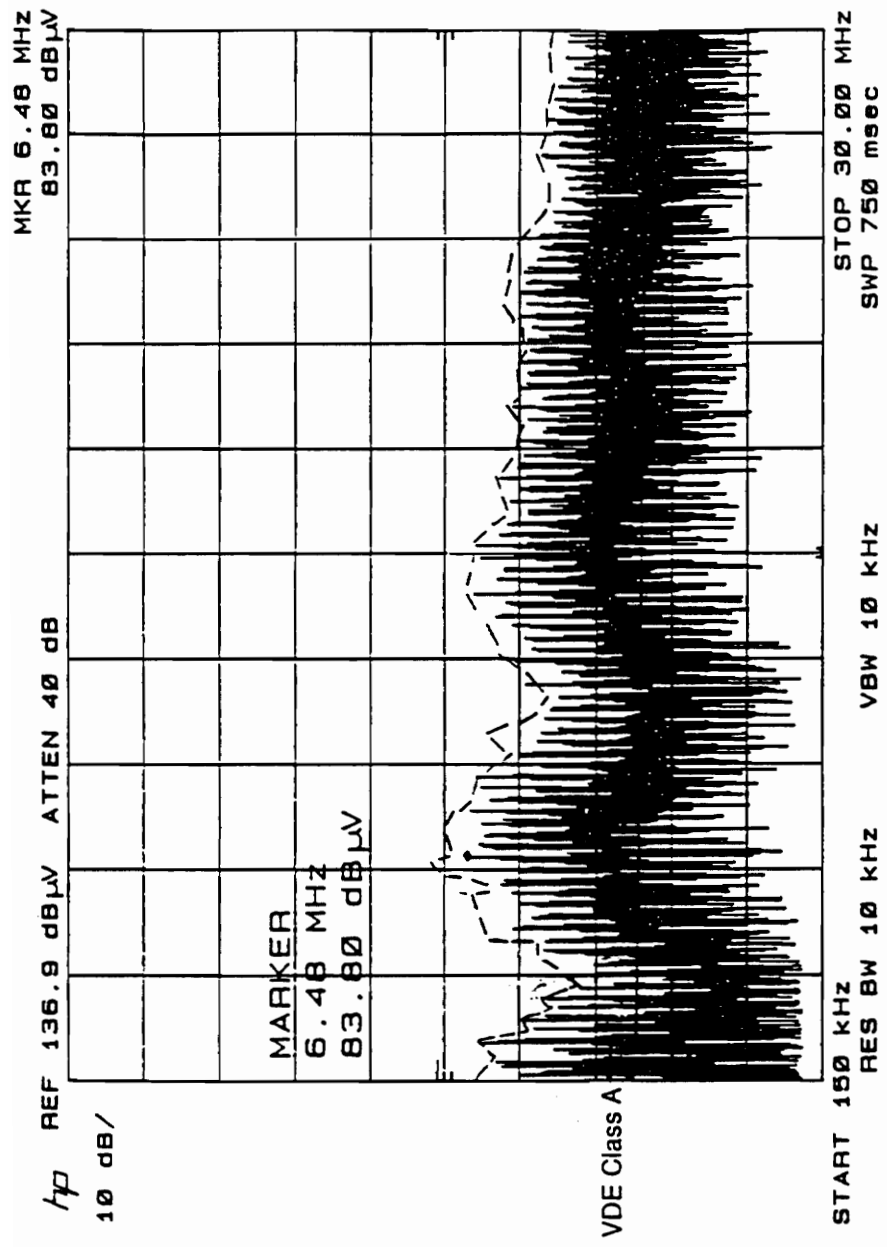
Figure 5.2.2 (b) Noise Spectrum Of A Forward Converter



Solid line: VDE Class A

Modulated,  $f_c = 90$  kHz, 50 kHz--150 kHz,  $\Delta f = 15$  kHz,  $f_m = 400$  Hz , RBW = 300 Hz

Figure 5.2.2 (c) Noise Spectrum Of A Forward Converter



Dotted line: envelop of unmodulated spectrum

Solid line: VDE Class A

Modulated,  $f_c = 90$  kHz, 150 kHz--30 MHz,  $\Delta f = 15$  kHz,  $f_m = 400$  Hz,  $RBW = 10$  kHz

Figure 5.2.2 (d) Noise Spectrum Of A Forward Converter

because 200 Hz and 9 kHz setting are not available from the analyzer model used. Figures 5.2.2 ( c ) and 5.2.2 ( d ) show the results when frequency modulation scheme is used. The carrier frequency  $f_c$  ( PWM frequency of power supply ) is approximately 90 kHz.  $F_m$  of 400 Hz and  $\Delta f$  of 15 kHz were used. It is evident from Figure 5.2.2 ( c ) that side-band harmonics of the fundamental switching frequency are created. The amplitude envelope of the unmodulated test is indicated by the dotted line, and the VDE Class A specification is indicated by the solid line. Improvement due to FM is clearly shown, especially at fundamental switching frequency. Without using FM, the emission cannot pass VDE class A at the fundamental frequency. With the FM, it does. From 150 kHz to 30 MHz, the improvement is not as pronounced, and the emission cannot pass VDE class A specification even with modulation. However, the emission at fundamental frequency dictates the size of EMI filter, therefore, EMI filter can still be smaller by using FM scheme.

### 5.3 Parameters Affecting Performance

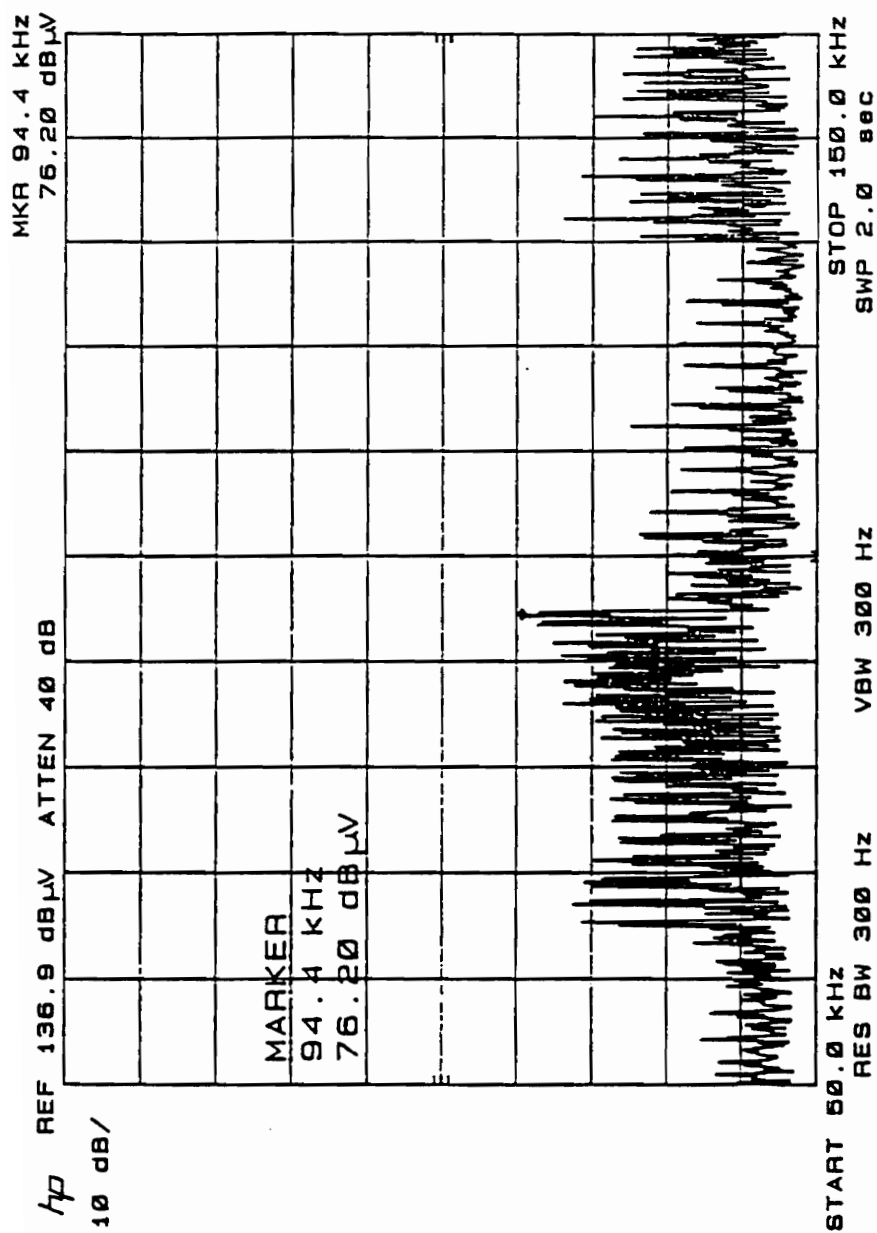
There are several parameters that affect the modulating effect as described in Chapter IV. Mathematically speaking,  $\beta$  (  $= \Delta f / f_m$  ) value alone determines the resultant frequency spectrum. In practice, however, the measurement requirement dictated by FCC or VDE rules also affect the result.

#### 5.3.1 $\beta$ Value

In theory, the larger the  $\beta$  value, the more even is the resultant spectrum. This is evident from Figure 4.2.1. However, because of measurement bandwidth requirements,

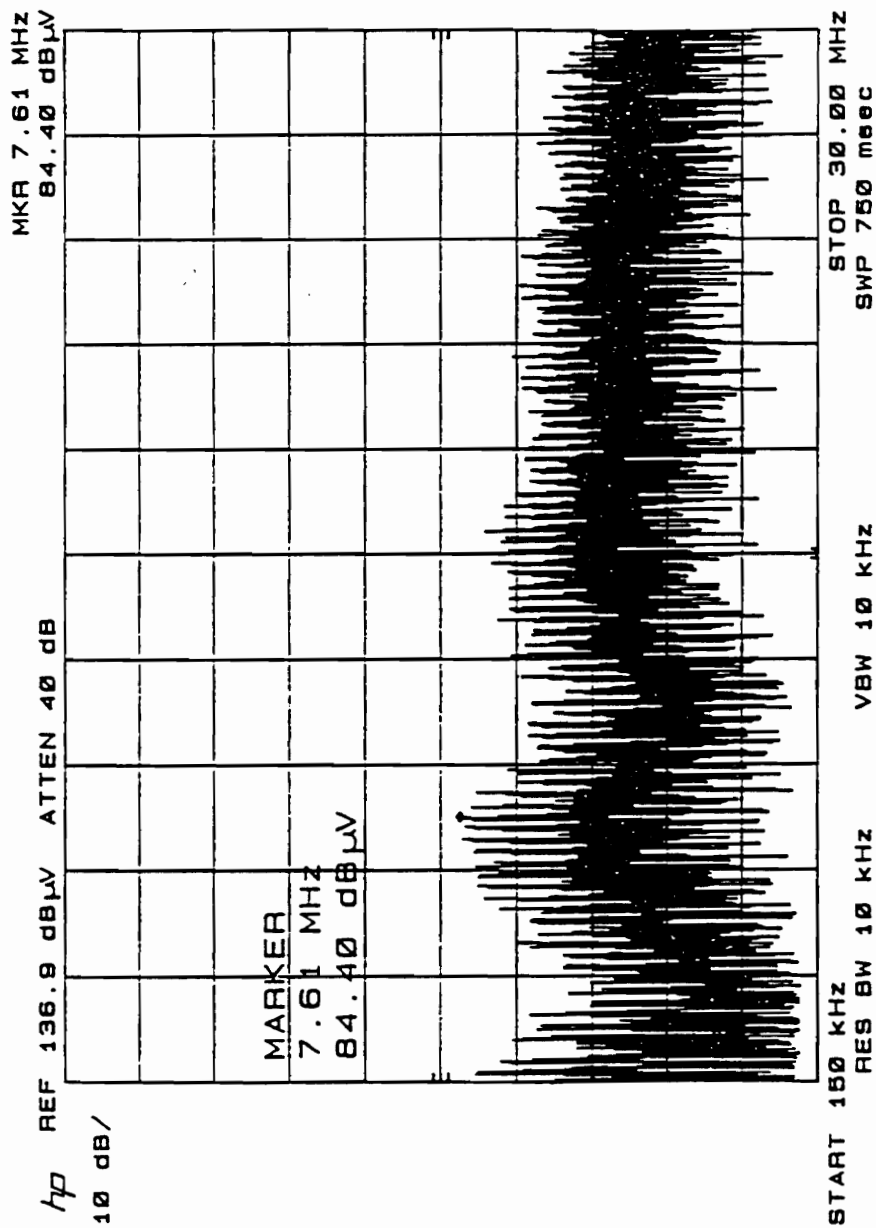
increasing  $\beta$  may not result in any significant reduction in the measured EMI. For example, if  $f_m$  is significantly smaller than the measurement bandwidth requirements ( 200 Hz for  $10 \text{ kHz} < f < 150 \text{ kHz}$ , and 9 kHz for  $f > 150 \text{ kHz}$  ), then the side-band harmonics are clustered together, and the measured result will not show any significant improvement by using FM, even though mathematically it should. To the spectrum analyzer set at relatively large RBW, the side-band harmonics are indistinguishable, and the measurement result shows the total power covered in the RBW as one large harmonic. Carson's rule described in Section 4.3 states that the total power contained in a FM signal does not change, therefore, the measured result will not be changed by small  $f_m$ . Figures 5.2.2 ( c ), 5.3.1.1 ( a ) and 5.3.1.2 ( a ) show the effect of changing  $\beta$  value by changing  $f_m$  value.  $\Delta$  value is kept the same at 15 kHz in all these three figures.  $f_m$  is, respectively, 400 Hz, 200 Hz and 600 Hz. It can be seen that the modulation effect in part ( c ) or ( a ) ( 10 kHz -- 150 kHz ) of each of the three figures is very evident. It is also seen that the noise reduction effect of FM is most effective when  $f_m = 400 \text{ Hz}$ , even though  $\beta$  value is not the largest among the these cases involved. For part ( d ) or ( b ) ( 150 kHz -- 30 MHz ) of all of these figures, the effect of FM is observable, but the difference between each case is minimal. This will be explained in Section 5.3.2.

To be effective in EMI reduction, therefore,  $f_m$  should be at least close to or larger than the required RBW. Take Figure 5.2.2 ( c ) as an example,  $f_m$  is 400 Hz, and the effect of the shattering of the fundamental frequency harmonics is very evident, because RBW is only 300 Hz. Each two adjacent side-band harmonic is separated by the space larger than the RBW. For measurement frequency higher than 150 kHz, RBW is set at 10 kHz, the effect of FM shattering may or may not be significant depending on the frequency range.



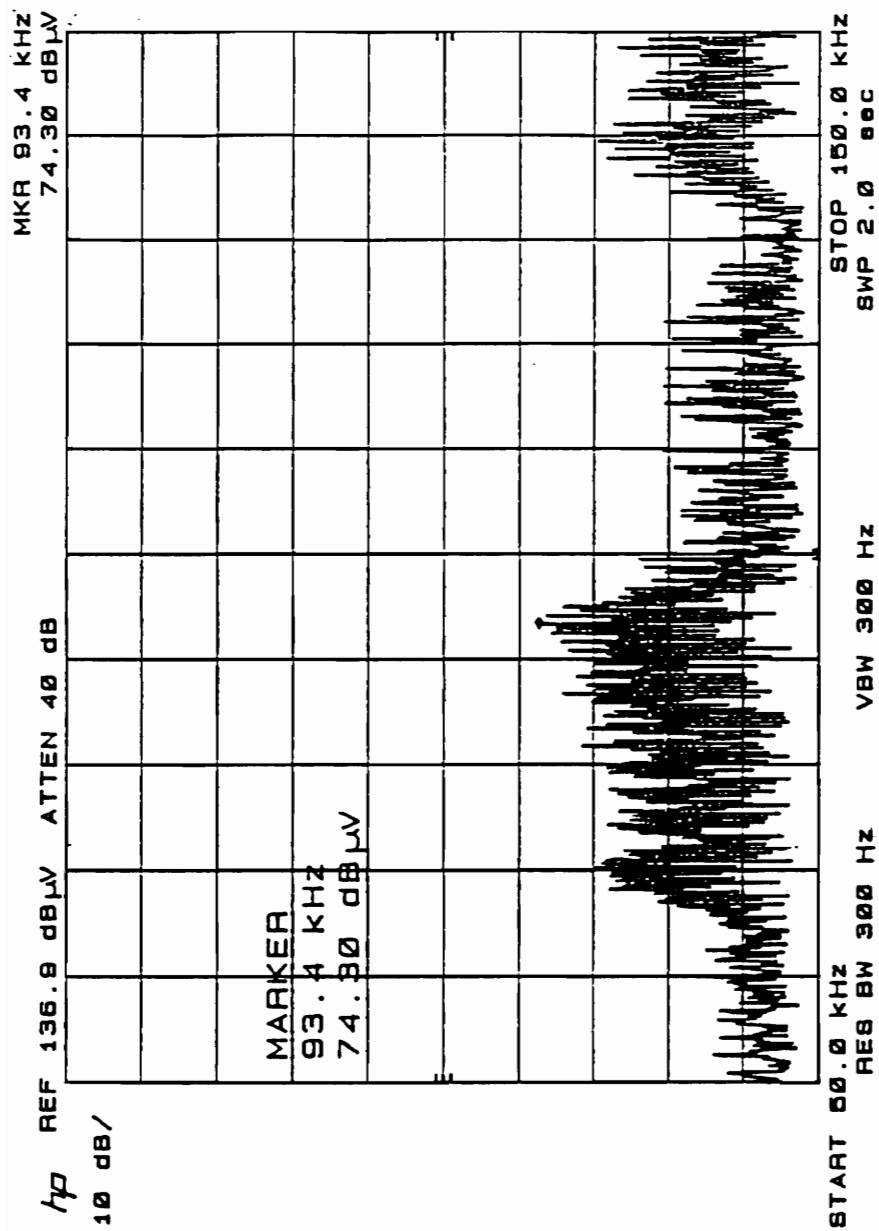
Modulated,  $f_c = 90$  kHz,  $50$  kHz-- $150$  kHz,  $\Delta f = 15$  kHz,  $f_m = 200$  Hz,  $RBW = 300$  Hz

Figure 5.3.1.1 (a) Noise Spectrum Of A Forward Converter



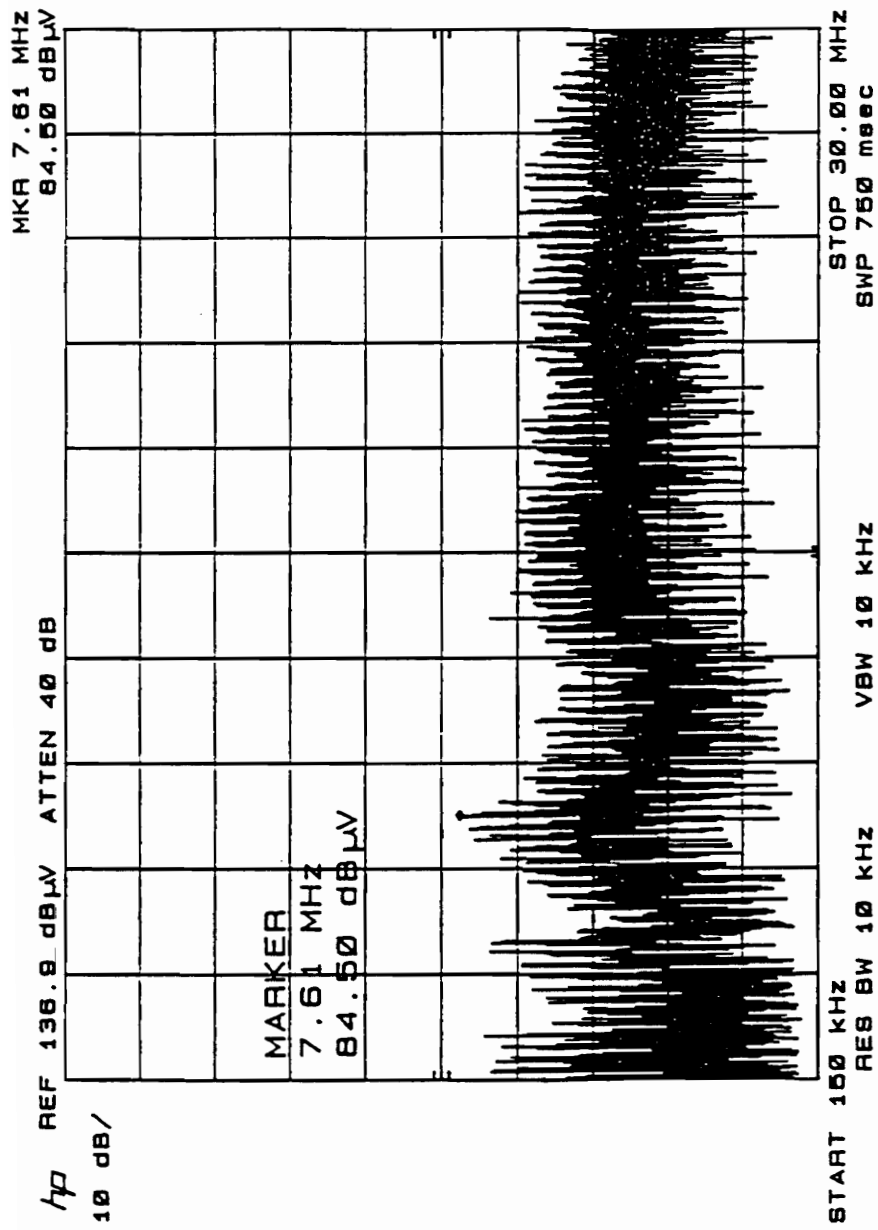
Modulated,  $f_c = 90 \text{ kHz}$ ,  $150 \text{ kHz}$ --  $30 \text{ MHz}$ ,  $\Delta f = 15 \text{ kHz}$ ,  $f_m = 200 \text{ Hz}$ ,  $\text{RBW} = 10 \text{ kHz}$

Figure 5.3.1.1 (b) Noise Spectrum Of A Forward Converter



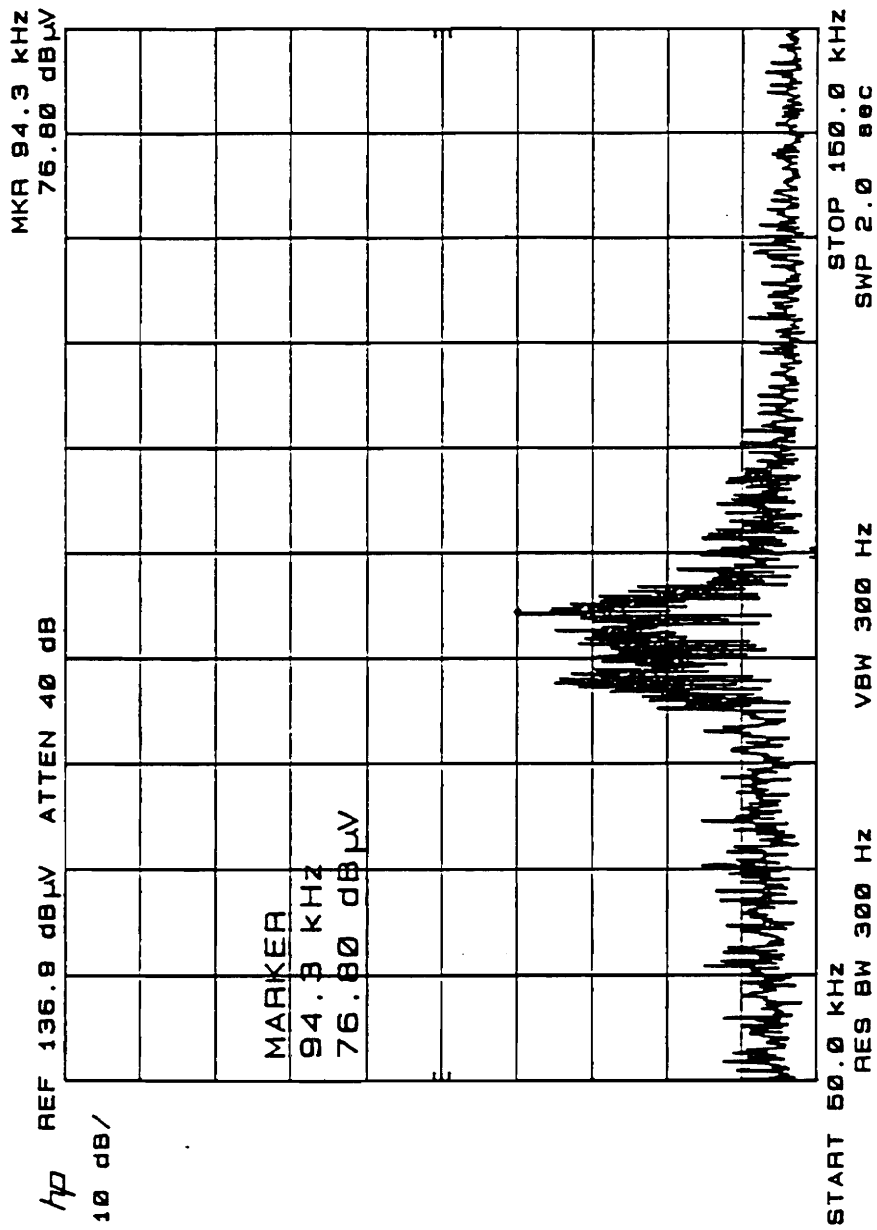
Modulated,  $f_c = 90$  KHz,  $50$  KHz-- $150$  KHz,  $\Delta f = 15$  KHz,  $f_m = 600$  Hz,  $RBW = 300$  Hz

Figure 5.3.1.2 (a) Noise Spectrum Of A Forward Converter



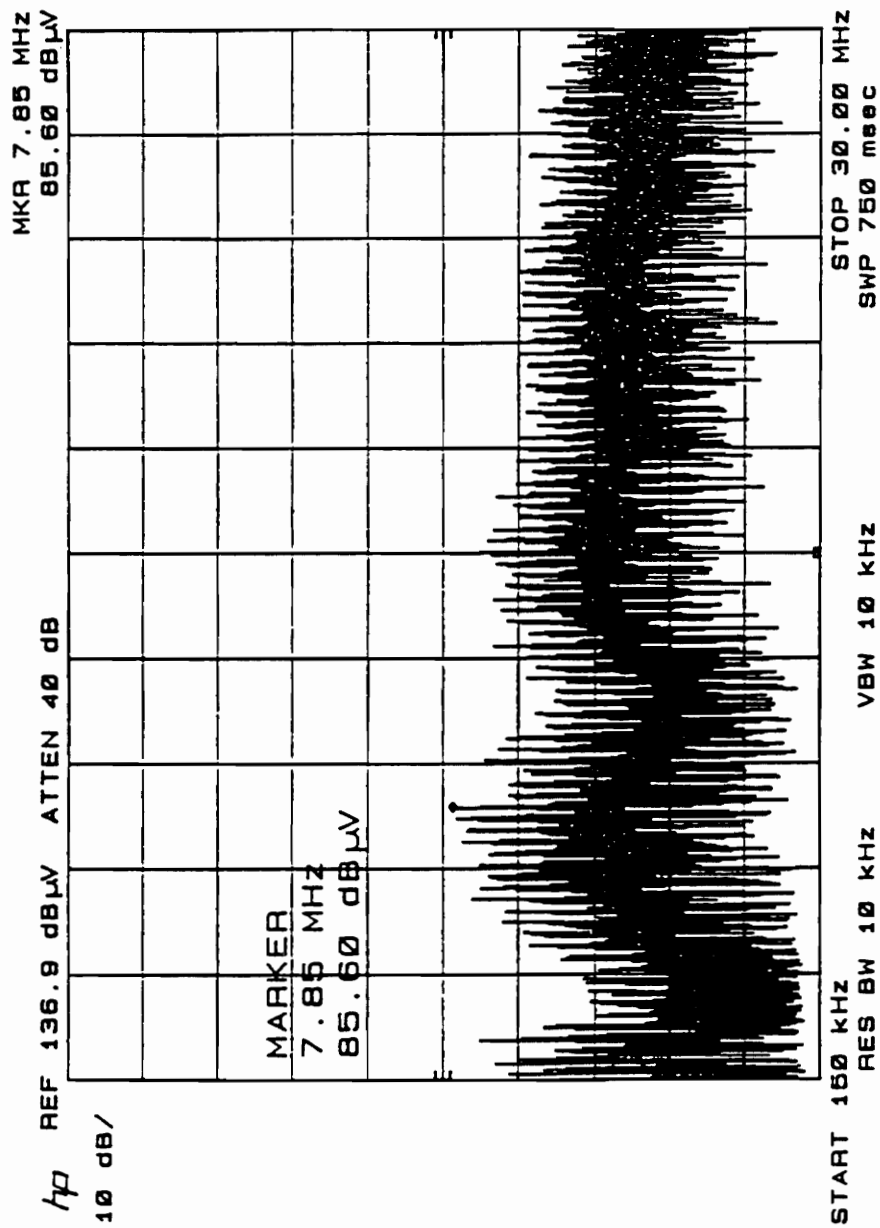
Modulated,  $f_c = 90 \text{ kHz}$ ,  $150 \text{ kHz} \rightarrow 30 \text{ MHz}$ ,  $\Delta f = 15 \text{ kHz}$ ,  $f_m = 600 \text{ Hz}$ ,  $\text{RBW} = 10 \text{ kHz}$

Figure 5.3.1.2 (b) Noise Spectrum Of A Forward Converter



Modulated,  $f_c = 90$  kHz, 50 kHz--150 kHz,  $\Delta f = 7.5$  kHz,  $f_m = 400$  Hz,  $RBW = 300$  Hz

Figure 5.3.2 (a) Noise Spectrum Of A Forward Converter



Modulated,  $f_c = 90$  kHz, 150 kHz--30 MHz,  $\Delta f = 7.5$  kHz,  $f_m = 400$  Hz, RBW = 10 kHz

Figure 5.3.2 (b) Noise Spectrum Of A Forward Converter

### **5.3.2 $\Delta f$ Value**

From the EMI reduction point of view, larger  $\Delta f$  value should be selected because it leads to larger  $\beta$  value. Figures 5.2.2 ( c ) and 5.3.2 ( a ) show the difference in FM effect when  $\Delta f$  is changed.  $f_m$  is 400 Hz for both cases. The FM effect in Figure 5.3.2 ( a ) is less because  $\beta$  value is smaller due to smaller  $\Delta f$  value. However, larger  $\Delta f$  leads to wider frequency variation which may not be acceptable in some applications. Practically, side-band harmonic frequency should be at least above human hearing range.

High  $\Delta f$  has another effect on the measured EMI. According to Carson's rule, the side-band bandwidth  $B_{nT} \approx nB_T$  for each switching harmonic. For higher order switching harmonics ( large  $n$  ) and large  $\Delta f$ , the side-band bandwidth is large. Therefore, side-band harmonics of different switching-frequency harmonics may overlap. Part ( b ) or ( d ) of all the figures are not much different from each other. Different  $f_m$  values or  $\beta$  values lead to different theoretical spectra, but to the spectrum analyzer with 10 kHz RBW, they all look the same, because  $f_m$  is much smaller than RBW, and the side-band harmonics overlap. Unless  $n\Delta f \gg \text{RBW}$  and  $\Delta f/f_m \gg 1$ , the FM effect will not be pronounced.

### **5.3.3 $f_m$ Value**

The effect of  $f_m$  value on EMI is two fold. On one hand,  $f_m$  should be small to have large  $\beta$  value. On the other hand,  $f_m$  should be large compared to spectrum analyzer RBW, in order to have significant noise reduction effect. Also,  $f_m$  value should be small to avoid disturbing the power supply output voltage regulation. Thus,  $f_m$  should be significantly less than the control loop gain crossover frequency. Figure 5.3.1.1 ( a ) shows the effect when  $f_m = 200$  Hz. Because  $f_m$  is smaller, the frequency components are closer to

each other, the effect is worse than when  $f_m$  is 400 Hz.

#### **5.3.4 Table Summarizing The Effect Of Parameter Change**

Table 5.3.4 summarizes the effect of varying  $f_m$  and  $\Delta f$  on noise reduction. When  $\beta$  value is increased, the noise reduction should be better theoretically. But, due to the RBW requirement, the result may or may not be improved.

### **5.4 Possible Adverse Effects Caused By FM Of PWM Frequency**

#### **Resonance With EMI Filter**

In a power supply circuit, there is a possibility of parasitic resonance between an EMI filter and the source impedance. The emission noise could be amplified by the filter at the parasitic resonance frequency. In using an FM scheme, the harmonics spread cover a wide range. This increases the chance of hitting the parasitic resonance frequency and this scheme may actually degrade the emission performance at the resonance frequency. To prevent this from happening, the filter must be properly damped to reduce the Q of the parasitic resonance.

#### **Output Regulation**

The bandwidth of the control loop gain of a power supply is limited. If  $f_m$  exceeds the control bandwidth, then the output voltage regulation will be adversely affected.  $F_m$  should therefore be selected to be less than the loop gain bandwidth.

### **Noise In The Forbidden-Band**

Since side-band harmonics are generated in a FM approach, it must be ensured that no significant harmonics lie in the human hearing frequency range.  $\Delta f$  should be chosen small enough to avoid this potential problem. For some applications, certain frequency band is forbidden. It should be made sure that the side-band harmonics do not spread into the forbidden zone.

Table 5.3.4 Effect Of Varying  $F_m$  and  $\Delta f$  On Noise Reduction

$f_m$	$\Delta f$	$\beta = \Delta f / f_m$	Result
same	smaller	smaller	worse
smaller	same	larger	if $f_m > \text{RBW}$ , better if $f_m < \text{RBW}$ , could be better or worse

## Chapter VI

### CONCLUSIONS AND FUTURE RESEARCH

#### 6.1 Conclusions

There are several conclusions drawn from the investigation:

- (1) Conducted EMI of a switching power circuit can be reduced by modulation of the PWM switching frequency. This possibility has been verified experimentally.
- (2) The modulating parameters must be properly chosen to have effective EMI reduction without causing significant side effects, such as audible noise due to frequency variation, and poor converter output voltage regulation. Reduction of more than 10 dB is practical.
- (3) Due to the required resolution bandwidth ( RBW ) of the measuring spectrum analyzer, frequency modulation is particularly effective in reducing the fundamental harmonic, if the switching frequency is less than 150 kHz. Modulating frequency  $f_m$  should be chosen to be somewhat greater than the required RBW of 200 Hz. This choice is effective in reducing EMI fundamental harmonic without causing many adverse side effects. If the power supply switching is above 150 kHz, then a much larger modulating frequency should be chosen ( 9 kHz ) to be effective in reducing the fundamental noise component. Such a choice in some cases may cause significant adverse side effects such as converter output voltage regulation. A compromise of using lower modulating frequency may be necessary.

- (4) Both the common-mode emission and the differential-mode emission can be reduced by frequency modulation. However, the emission generated by ringing can not be reduced by this approach. Snubber circuit and use of soft diodes alleviate this problem.
- (5) Since the side-band harmonics generated by frequency modulation are scattered over a wide frequency range, the EMI filter must be properly damped to avoid possible amplification of emission at the filter pole frequencies.
- (6) The resonant class of converters, in general, are operated with variable switching frequency over the input voltage and the output power range. However, if the input voltage and the load power are fixed during EMI measurement, then the switching frequency is constant, and there is no modulation effect. If the resonant converter is operated off-line, then the DC input voltage contains 120 Hz ripple, which causes modulation of switching frequency; EMI emission reduction is then possible.

## 6.2 Future Research

- (1) The effect of frequency modulation on the radiated EMI should be similar to that of the conducted EMI, even though this possibility is not experimentally verified. Radiated EMI specification ranges from 30 MHz to 1 GHz, and the required measurement resolution bandwidth ( RBW ) is 120 kHz. This may seem that large  $f_m$  is required for effective reduction. However, the side-band harmonics of the very high order switching-frequency harmonics ( 30 MHz -- 1 GHz ) cover large bandwidth (  $= 2n\Delta f$  ), and the emission can be effectively reduced as long as  $2n\Delta f > 120$  kHz. Therefore, with moderate modulating signal frequency, noise reduction in radiated EMI can be achieved. But this required further research to confirm experimentally.
- (2) Single-frequency modulating signal is used in the analysis of spectrum and in the

experiment. Other forms of modulating signal are possible, including a randomly modulated carrier frequency [11]. Further research is needed to confirm this experimentally and compare the difference in the effectiveness of different modulation scheme.

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